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PROCEEDINGS OF

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Volume 16 January, 1928 Number 1

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GENERAL INFORMATION

The PROCEEDINGS of the Institute are published monthly and contain the papers and the discussions thereon as presented at meetings.

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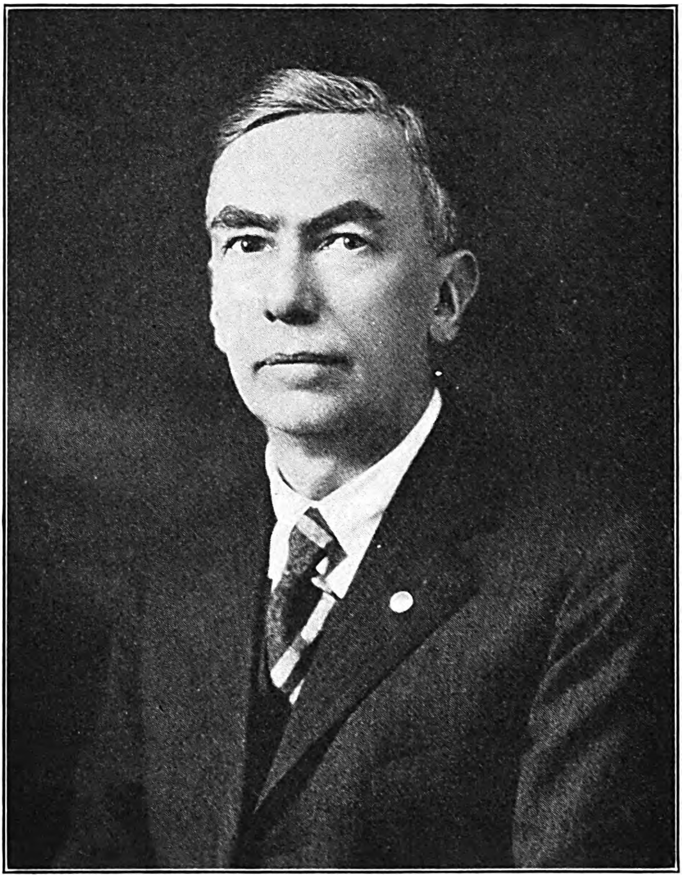


Frank Conrad
Vice-President, Institute of Radio Engineers—1927

Frank Conrad

VICE-PRESIDENT OF THE INSTITUTE, 1927

Frank Conrad was born in 1874 in Pittsburgh, Pennsylvania. In 1890 he entered the employ of the Westinghouse Electric and Manufacturing Company in Pittsburgh as an assistant in the shops making registering trains for the Shallenberger ampere-hour meters. Mr. Conrad's rise in the Westinghouse organization was rapid. He entered the laboratory after several years as an assistant in the shops. During this stage of his work he invented a number of forms of switches, lightning arresters, and breakers for use in alternating current work. He was closely associated with, and later in entire charge of the Arc Lamp Design Department. This was his first engineering work. Mr. Conrad's connection with radio dates back before the days of any broadcasting. He became interested as an amateur in radio reception of time signals and later in radio-telephone transmission by means of vacuum tubes. He established an amateur radio telephone station which later resulted in the development of the Westinghouse Station, KDKA.

Mr. Conrad was appointed a General Engineer of the Westinghouse organization in 1904 and was promoted to the position of Assistant Chief Engineer in 1921.

In 1925, he was awarded the Morris Liebmann Memorial Prize of the Institute for his early work in connection with high-frequency transmission. He is a Fellow of the Institute of Radio Engineers, a Member of the American Institute of Electrical Engineers, a member of the Society of Automotive Engineers, and a member of the American Association for the Advancement of Science. Mr. Conrad was elected Vice-President of the Institute in 1927 and was the Chairman of the Committee on Admissions during 1927.

CONTRIBUTORS TO THIS ISSUE

Armstrong, Edwin H., Born December 18th, 1890. Was educated at Columbia University, E. E. degree. Research graduate under Professor M. I. Pupin in the Hartley Research Laboratory at Columbia University for a number of years. During the war a captain and later major in the Signal Corps of the United States Army in the Army Research Laboratories in Paris, France. Mr. Armstrong is universally known for his work in connection with the regenerative circuit, the superheterodyne circuits and the superregenerative circuit. At present he is doing research work in the Hartley Research Laboratory of Columbia University. In 1918 Mr. Armstrong was awarded the Institute Medal of Honor. He has been a frequent contributor to the PROCEEDINGS of the Institute, and is a Fellow of the Institute.

Hazel, Herbert C.: Born at Harrodsburg, Indiana, September 26th, 1899. Received A.B. degree Indiana University 1922; received M.A. degree, 1926. During research connected with this work on problems of measuring current and resistance at radio frequencies, the thermionic vacuum tube method of calibrating ammeters was devised. At present, critic teacher in physics in Bloomington (Indiana) High School. Member of Phi Beta Kappa and of Indiana Academy of Science.

Heising, Raymond A.: Born at Albert Lea, Minnesota, August 10th, 1888. Received the E. E. degree, University of North Dakota, 1912; M.S. degree, University of Wisconsin, 1914. With the Research Department, Radio Section, Western Electric Company since 1914 until the Bell Telephone Laboratories were organized. At present with the latter concern. Connected with the development of transmitting apparatus, designed and operated the Arlington Trans-Atlantic Telephone Transmitter, 1915; invented constant current modulation system used extensively in commercial broadcast work and by the United States Army and Navy. Since the war continued research and development work in connection with ship to shore operation, trans-atlantic tests, etc. Mr. Heising is a Fellow of the Institute. He has contributed numerous papers to the PROCEEDINGS of the Institute and is a member of its Board of Direction. In 1921, he was the recipient of the Institute's Morris Liebmann Memorial Prize.

Marconi, Guglielmo.: Born at Bologna, Italy, April 25th, 1874. Educated at Leghorn Technical School. In 1895 began series of experiments on communication by means of Hertzian waves at which time he was able to transmit intelligible signals to a distance of 1½ miles. In 1896 he took out the first patent granted for a practical system of wireless telegraphy. He transmitted the first signals across the Atlantic in 1901. Senatore Marconi has been associated constantly with the development of radio telegraphy and telephony through his connection with the British Marconi Company and its associated companies. In 1920 Senatore Marconi was awarded the Institute Medal of Honor He has been a frequent contributor to the PROCEEDINGS of The Institute of Radio Engineers and is a Fellow of the Institute.

Wheeler, Harold A.: Born May 10th, 1903. Assistant in tests of radio receiving equipment, Bureau of Standards, 1921–1922. Associated with Professor Hazeltine in the study of neutralization of capacity coupling in vacuum tubes, 1922–1923. Engineering, Hazeltine Corporation in development of Neutrodyne receivers, including automatic volume control, 1924 to present. Received the B.S. degree in physics, George Washington University, 1925. Assistant in Physics Department, Johns Hopkins University, 1925 to date. Associate member of the Institute.

INSTITUTE ACTIVITIES

1928 Convention Plans

PLANS for the 1928 Convention have been completed. The Convention will be held on January 9th, 10th, and 11th, the Convention Headquarters being in the lobby of the Engineering Societies Building, 33 West 39th Street, New York City.

Papers on the following subjects will be presented: "A Digest of The International Radiotelegraph Conference;" several papers on "Audio Frequency Amplifiers"; "The Making of Talking Moving Pictures" (with demonstration); "Radio Picture Transmission Symposium on Inter-Electrode Tube Capacities," and several others.

The inspection trips will include a bus ride through the new Holland Tunnel, through the Experimental High Power Station group at Whipanny, New Jersey, and the National Broadcasting Company's Station, WJZ; an inspection of the technical equipment of Roxy's Theater; inspection of the studios of the National Broadcasting Company; inspection of the F. A. D. Andrea plant; inspection of Aerovox plant. The plans call for a dinner with entertainment on the evening of January 11th.

December Meeting of the Board of Direction

At the meeting of the Board of Direction held on December 7th in the offices of the Institute the following were present:

Ralph Bown, President; A. N. Goldsmith, Secretary; Melville Eastham, Treasurer; L. A. Hazeltine, R. A. Heising, R. H. Marriott, R. H. Manson, Donald McNicol, and J. M. Clayton, Assistant Secretary.

The following were transferred or elected to higher grades in the Institute:

Transferred to the grade of Member; T. G. Deiler, B. A. Engholm, W. H. Fortington, L. J. Gallo, J. J. Stanley and John A. Victoreen.

Elected to the grade of Member; E. A. Beane, L. B. Root, and J. K. Skirrow.

One hundred and thirty eight Associate and eight Junior members were elected.

Bound Volumes

Due to the increase in the number of pages in the PROCEEDINGS throughout the year it has been necessary to increase the price of the 1927 *Bound* Volumes. These volumes in blue buckram can

be purchased for $12.00 in the United States and Canada and $13.00 in other countries. Members of the Institute, libraries, and book sellers are entitled to a 25 percent discount from these prices.

Bound volumes are available from 1917 to 1927 inclusive. The cost of those prior to 1927 is $8.75 to members of the Institute, book dealers, and libraries, and $11.00 to non-members.

Institute Meetings

New York Meeting

At the New York Meeting of the Institute held on December 7th in the Assembly Room of the Engineering Societies Building, 33 West 39th Street, R. A. Heising of the Bell Telephone Laboratories presented a paper, "Experiments and Observations Concerning the Ionized Regions of the Atmosphere."

Following the presentation of the paper the following participated in its discussion: Messrs. Ballantine, Hallborg, Ohl, Rybner, Shaughnessy, and Brandt.

Over three hundred members attended the meeting.

A limited number of preprint copies of the paper in pamphlet form are available free of charge to members of the Institute upon application to the Institute office.

Atlanta Section

A meeting of the Atlanta Section was held on December 7th in Room 207 of the Chamber of Commerce Building, Atlanta, Georgia. Major Walter Van Nostrand was the presiding officer. C. F. Daugherty read Major Armstrong's paper on "Method of Reducing the Effect of Atmospheric Disturbances." An informal discussion followed the reading of the paper.

The next meeting of the Atlanta Section will be held on January 4, 1928.

Buffalo-Niagara Section

At the meeting of the Buffalo-Niagara Section held on November 17th in Foster Hall of the University of Buffalo Dr. Ralph Bown, President of the Institute, presented a paper on "Trans-Atlantic Telephone." L. C. F. Horle was the presiding officer.

A general discussion followed the presentation of this paper.

The attendance at this meeting was over 130.

On December 2d the Buffalo-Niagara members attended a meeting of the American Institute of Electrical Engineers to hear two papers, one by C. H. Bell of the Gould Storage Battery

Company on "Storage Battery Engineering and Development," and the other by A. T. Hinckly of the U. S. Light and Heat Corporation entitled "Dry Battery Developments."

Chicago Section

There will be a meeting of the Chicago Section on December 16th. Dr. Fred W. Kranz of Riverbank Laboratories will read a paper on "Some Characteristics of Speech and Hearing."

Connecticut Valley Section

The Connecticut Valley Section held a meeting on December 2d in the club rooms of the United Electric Light Company of Springfield, Massachusetts. Dr. W. G. Cady was the presiding officer.

Carl J. Madsen delivered a paper on "Recent Experience in Installing a 20 Kilowatt Radio Transmitter in Siberia." A general discussion followed the presentation of this paper.

For 1928 the following officers of the Section were elected:

Chairman, Dr. W. G. Cady; Vice-Chairman, Dr. K. S. Van Dyke; Secretary-Treasurer, George W. Pettingill.

Cleveland Section

On December 2d a meeting of the Cleveland Section was held in the Case School of Applied Science. John R. Martin presided.

C. A. Wright presented a paper entitled "Measurement of Inductance." F. T. Bowditch read a paper on "The Measurement of Choke Coil Inductance." Messrs. Victoreen, David, Martin and others discussed these papers.

The new officers of the Cleveland Section were elected as follows: Chairman, J. R. Martin; Vice-Chairman, D. Schregardus; Secretary-Treasurer, B. W. David.

Ralph Farnham was appointed Chairman of the Meetings and Papers Committee and John A. Victoreen was appointed Chairman of the Publicity Committee.

Forty-four members attended this meeting.

Detroit Section

A meeting of the Detroit Section was held on November 18th in the Conference Room of the Detroit News Building, Detroit, Michigan. Thomas E. Clark was the presiding officer.

Stanley R. Manning of the Engineering Department of the Michigan Bell Telephone Company read a paper on "The Effect of Filter Circuits Upon Audio Frequencies." There was included

a demonstration of the elimination of frequencies through filters, specially prepared phonograph records being used.

Thirty-two members of the Section were present at this meeting.

On December 16th there will be a meeting of the Section in the Michigan Bell Telephone Building at which time Dr. N. H. Williams will deliver a paper on "Some Operating Characteristics of The Screen Grid Vacuum Tube." It is planned that a banquet will be held before the next meeting.

Philadelphia Section

A meeting of the Philadelphia Section was held in the Franklin Institute on November 25th. J. C. Van Horn presided.

Dr. J. P. Maxfield delivered a paper on "Electrical Recording and Reproducing."

Two hundred and seventy five members of the Section and their guests attended this meeting.

The next meeting of the Section will be in the Bartol Laboratories on January 27th.

San Francisco Section

Plans for the reorganization of the San Francisco Section are actively under way. A meeting of a committee of the Section membership was held in November. A general meeting, for the purpose of electing officers and making business arrangements for the resumption of the Section's activities, will be held on December 29th. It is expected that regular meetings of the San Francisco Section will be resumed during January of 1928.

Prominent in the reorganization work are: Dr. Leonard F. Fuller (Member Committee on Sections), D. B. McGown, B. H. Linden, A. Y. Tuel and others.

Seattle Section

The October meeting of the Seattle Section was held on the 15th of the month in the Club Rooms of the Telephone Building, Third and Seneca Streets, Seattle, Washington, T. M. Libby presided.

Two papers were presented, the first by F. A. Brown was entitled, "Inspection of Fishers Blend Station KOMO." The second paper was presented by W. A. Kleist and was entitled, "Trans-Atlantic Telephone Circuit."

Messrs. Libby, Rowe, Renfro, Deardorff, Sylvester, and others discussed these papers.

Proposed Pittsburgh Section

There has been received at headquarters of the Institute a petition from members residing within the vicinity of the Pittsburgh territory asking for approval of the formation of a Pittsburgh Section of the Institute. W. K. Thomas, of Ludwig Hommel, has been the leader in the organization of this territory. This petition will be presented to the Board of Direction at the January meeting thereof.

Committee Work

Committee on Admissions

At the meeting of the Committee on Admissions held in the offices of the Institute on December 7th were the following members: L. A. Hazeltine, F. K. Vreeland, and H. F. Dart.

Eighteen applications for transfer or admission to the higher grades of membership in the Institute were considered. The committee approved the applications for transfer to the grade of Member from six associates. The committee also recommended that three applicants be directly elected to the Member grade.

Committee on Sections

At a meeting of the Committee on Sections, October 26th, Messrs. Gage (Chairman), Berger, Shute, and Clayton were present.

The Committee reviewed correspondence with members in the San Francisco territory regarding the reorganization of the San Francisco Section.

Plans for a Section representatives' meeting at the 1928 Convention were formulated.

Committee on Membership

On October 31st a meeting of the Committee on Membership was held in the offices of the Institute. Messrs. Dart (Chairman), Finch, Shute, Brick, and Clayton were present.

The committee has been working for some time on a list of persons who are probably qualified for transfer to the higher grades of membership in the Institute.

The committee is considering comments whereby the Institute can be brought more closely in contact with students in engineering courses in technical schools throughout the country.

It was reported that the membership in the Institute during 1927 has undergone a *paid* increase of over forty-four per cent.

Patent Digests

Officers of the Institute have felt for some time that the Patent Digest Section of the PROCEEDINGS may not be serving the majority of the Members of the Institute to any great advantage. It is thought that these additional pages could be devoted to other matters with better advantage to the membership. Unless definite and concrete suggestions to the contrary are received in any quantity, the Patent Digest portion of the PROCEEDINGS will be eliminated immediately.

Personal Mention

W. C. Fogg has joined the Meter Testing Department of the Union Gas and Electric Company of Cincinnati, Ohio.

J. Kelly Johnson is now an instructor in electrical engineering, Columbia University, New York City.

C. B. Joliffee, formerly with the Buckeye Incubator Company of Springfield, Ohio, has rejoined the radio staff of the Bureau of Standards at Washington, D.C.

William F. Devine has become associated, as operating engineer, with Station WREN of Lawrence, Kansas. He was formerly in the Sales Department with the Broadway Radio Company of Kansas City.

Harry Diamond, formerly instructor in electrical engineering, Lehigh University, Bethlehem, Pennsylvania, is now an associate radio engineer at the Bureau of Standards, Washington, D.C. in the Radio Aeronautic Section in the Development of Radio Aids to Aviation.

Lewis M. Hull has recently become director of research of the General Radio Company of Cambridge, Massachusetts with headquarters at Cambridge. He continues to be the director of research of the Radio Frequency Laboratories of Boonton, New Jersey with which he has been associated for nearly six years. Through this new arrangement the facilities of the laboratories of both the General Radio Company and the Radio Frequency Laboratories will be available on problems of the other.

Errata

In the paper by Mr. Greenleaf W. Pickard on "The Relation of Radio Reception to Sunspot Position and Area," published in the PROCEEDINGS for December, 1927, the ordinates of Figs. 1 to 9, inclusive, have been erroneously designated as "Microvolts Per Meter." These ordinates represent percentage variation of the elements.

METHODS OF REDUCING THE EFFECT OF ATMOSPHERIC DISTURBANCES*

By

Edwin H. Armstrong

(Marcellus Hartley Research Laboratory, Columbia University)

Summary—*The transmitter sends the dots and dashes on one frequency and the spaces on a slightly different frequency. At the receiver a local frequency is superimposed producing two audio frequencies. The paths of the two audio frequencies are combined differentially so they would oppose if they occurred at the same time; also, they would pull the marking pen in the opposite direction if they occurred at the same time. Since one audio frequency is due to the dots and dashes while the other is due to the spaces, they do not occur at the same time. Since static hasn't a definite frequency it may produce about the same amount of audio-frequency current in each path, thereby more or less neutralizing its own effects. The tape records made with this system and an ordinary system show marked advantage in this system for reducing the effects of static and for increasing speed of recording.*

It is the purpose of this paper to describe a method of reducing the effects of atmospheric disturbances by selective means as distinguished from that means which depends on the geography of the situation: directional reception.

The method is based on the fact that the distribution of energy with respect to frequency of the waves of natural origin is such that over short periods of time the energy in the component at a given frequency is substantially equal to the energy at a closely adjacent frequency, and that in a "crash" or burst of static both these frequencies will be present simultaneously.

In the respect that the energy of the disturbances is distributed through a band and the energy of the signal concentrated in substantially one frequency there lies a fundamental difference which has been utilized to the utmost to effect the separation of the two by means of electrical tuning.

In the respect that the energy of the disturbances and the energy of the signal is alternately present or absent, the first by reason of the irregularities of nature, the second by reason (in telegraphy) of the Morse code or equivalent, there is no fundamental difference. In this respect the two are the same, the difference being one of degree only.

The method which is here described is based on the establishment of a difference between the natural waves and the signaling waves which lies in imparting to the signaling waves a charac-

* Received by the Institute, August 25, 1927.

* Presented before the Institute of Radio Engineers, New York City, October 5, 1927.

teristic not found in the waves of natural origin. This difference is established by producing at the transmitter two waves of closely adjacent frequency and radiating them alternately. Now consider that band of frequencies which just includes the two frequencies selected for signaling. Energy from the waves of natural origin will be received simultaneously in substantially equal amounts throughout the band. Energy from the signaling waves will be received alternately at the upper and lower ends of the band. With this fundamental difference it becomes possible

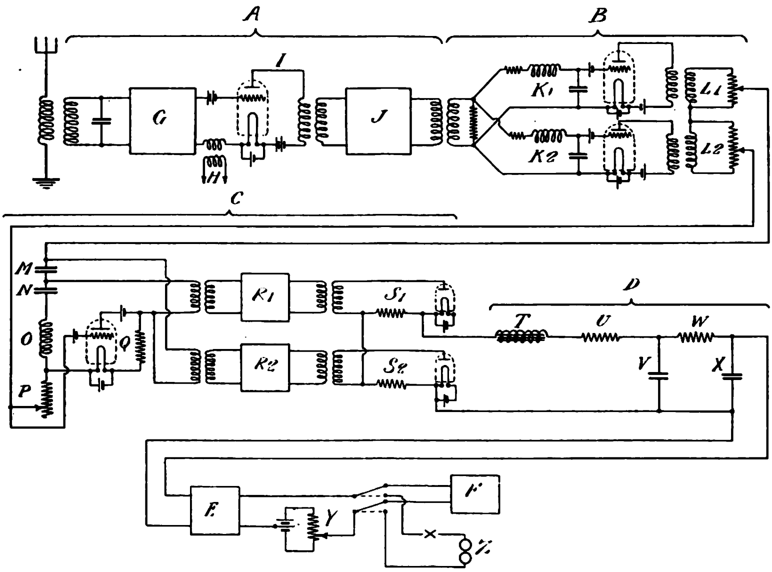


Fig. 1

to provide that long looked for device, the receiver which acts cumulatively with respect to the signal but differentially with respect to the "static." This receiver is one which produces a certain effect on the indicator when one of the two frequencies in question is present alone; an opposite effect on the indicator when the other frequency is present alone; but a neutralized or zero effect when both frequencies are present simultaneously.

Both the nature of the problem and the conditions of practical working require that the two signaling frequencies shall be very close to each other. Hence the method of selection must be capable of effectively separating the two frequencies in a minimum of time to meet the requirements of rapid signaling. In general,

on the long waves used in trans-oceanic signaling, a frequency variation of from 25 to 100 cycles is sufficient; on the shorter waves a greater variation (where the variation lies in the fundamental and not in a modulated or superimposed frequency) is necessary although the increase is not in any sense a proportional one.

The method of transmitting the two signaling frequencies requires no further comment than to say that the key is arranged to change the frequency transmitted when the key is down from, say 20,000 cycles, to 20,050 when the key is up.

The method of reception is illustrated by the arrangement of Fig. 1. In this figure A represents the usual form of receiver comprising a tuned amplifier G, a detector I, heterodyne H, and

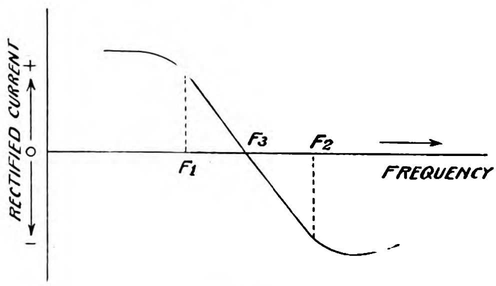


Fig. 2

low frequency amplifier J. Connected to the output of this low frequency amplifier are two tuned circuits K_1 and K_2 resonant respectively to the two beat frequencies composing the signal. In the case to which the curves hereinafter shown refer, these two frequencies were 1200 and 1280 cycles. Each circuit controls an amplifier and the outputs of the amplifiers are differentially connected through transformers and potentiometers L_1 and L_2. The organization just described has the double function of selectively responding to the two signaling frequencies with equal facility and of equalizing the energy of the natural waves between these two frequencies so that such irregularities as do occur are minimized.

The organization connected to the output of the equalizer and shown under the heading C represents the selective system for separating the two signaling frequencies, converting them to continuous currents and combining the resulting effects cumulatively. M, N, and O are condensers and an inductance whose values are so chosen that the combination of N and O is non-

reactive for 1220 cycles and the combination M, N, and O is non-reactive for 1280 cycles. The two combinations form the basis for the rapid separation of the two signaling frequencies. P, Q, and the associated vacuum tube amplifier is a resistance compensator for neutralizing the effect of the resistance in the condensers M and N, and of the inductance O. Q is a large resistance, normally four to five thousand ohms, and P is a resistance commensurate with the resistance of the combination M, N, O. Connected across the combinations N, O and the compensator and M N O and the compensator are two transformer primaries of

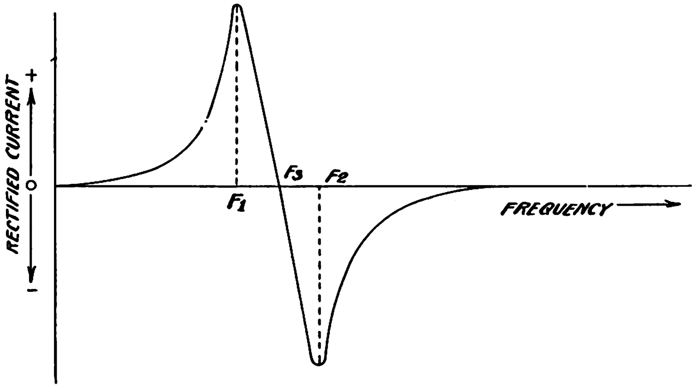


Fig. 3

sufficiently high impedance so that the characteristics of the system M N O are not affected. The secondaries of these transformers control two equal amplifiers R_1 and R_2 whose outputs are connected respectively to two valve rectifiers which are furnished with series resistances S_1 and S_2 of such value, that for the strength of signal used, substantially straight line rectification is effected. The outputs of these rectifiers is then differentially combined.

The relation between the combined continuous current output of the rectifiers and the alternating current input to the combination M N O with respect to frequency is shown in Fig. 2. As the frequency of the input current is increased the rectified current does not vary materially until a certain frequency F_1 is reached (1220 cycles). Further increase in frequency causes a decrease in the rectified current until it reaches zero at a certain frequency F_3. Still further increase of input frequency causes the output current to flow in the opposite direction and to increase until a certain frequency F_2 is reached beyond which the current again does not vary materially. The combined characteristic of the

equalizer B and selector C is illustrated in Fig. 3. Here the frequencies above F_2 and below F_1 are substantially cut off and a maximum and opposite response is obtained at each of these frequencies respectively.

Referring again to Fig. 1 the combined outputs of the rectifiers are fed into a low pass filter D as shown. The output of this filter is connected to a d-c. amplifier E whose output controls a siphon recorder F.

With this description of the system of Fig. 1 in mind examine the action of the arrangement for incoming signals (in the absence

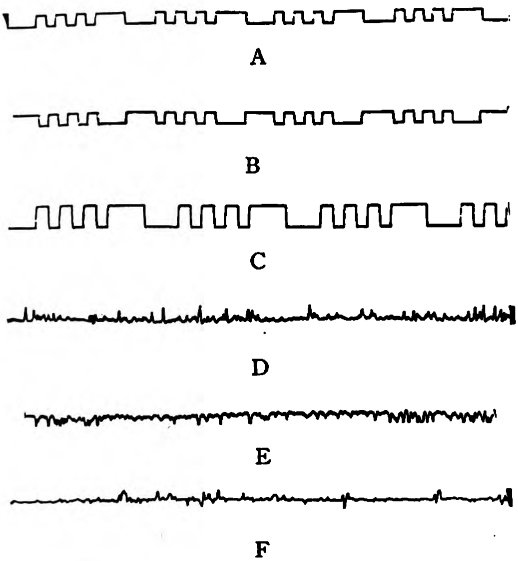
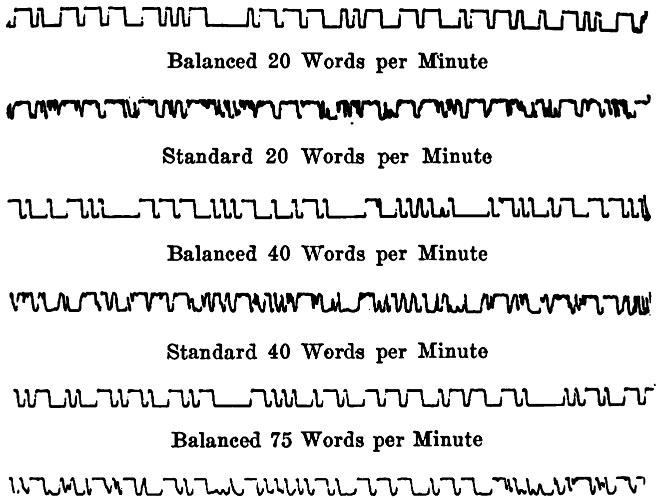


Fig. 4

of static): Suppose that the transmitted signal, with the key down, is 20,000 cycles and with the key up 20,060 cycles. Suppose the heterodyne to be adjusted to 18,780 cycles. Then when the key is up a beat frequency of 1280 cycles is produced and in accordance with the curve of Fig. 3 a rectified current having the polarity and magnitude corresponding to F_2 results. This passes through the filter D, is amplified by the d-c. amplifier E, and produces a deflection of the marker of the siphon recorder from the neutral position in a direction which depends on the polarity of the rectified current.

When the position of the transmitter key is reversed (i.e., down, 20,000 cycles transmitted) a beat frequency of 1220 cycles

is produced and this produces a rectified current corresponding in magnitude and direction to F_1 in Fig. 3. This rectified current is of substantially the same amplitude as that produced by the 20,060 cycles when the key was up, but it is of opposite polarity. This rectified current causes the marker of the siphon recorder to return to and pass through its neutral position and to be deflected in the direction opposite to that in which it was deflected when the transmitter key was up. Hence the total deflection of the siphon recorder is double that which would be obtained with the standard



Balanced 20 Words per Minute

Standard 20 Words per Minute

Balanced 40 Words per Minute

Standard 40 Words per Minute

Balanced 75 Words per Minute

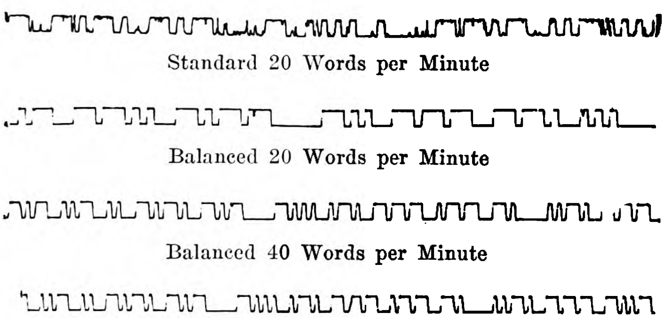
Standard 75 Words per Minute

Fig. 5

method of signaling. The three tapes shown in Fig. 4 illustrate this action. Tape *A* is a record in which only the 20,000 cycle current was transmitted, the letter *V* being sent by ordinary keying, (i.e., interruption). Tape *B* is the record produced when the same letter *V* is transmitted on 20,060 cycles, keying in the same way. In this case the response is inverted. Tape *C* is the record produced when keying is accomplished by alternately radiating both frequencies. In this case the response is substantially the cumulative combination of *A* and *B*.

Now consider the action of the system for the waves of natural origin (in the absence of signals). Over a period of time there will be substantially the same amounts of energy received from these waves between the frequencies F_1 and F_3 as between the fre-

quencies F_3 and F_2. Over short intervals of time due in part to the inequalities of heterodyning,* as well as to the irregularities of nature, more energy may be in the lower or higher of the two frequency bands. The equalizer *B* reduces this inequality so that there are delivered to the selector system *C* two bands of frequencies of about the same energy. These two bands are separated by the selector *C* into high and low frequency groups which are supplied respectively to the rectifiers. The character of the currents produced in the output of each of these rectifiers is illustrated in Fig. 4 by tapes *D* and *E*. Each current is a pulsating unidirectional one. When the two are combined differentially the resultant current shown in tape *F* is produced. It will be observed that this resultant current is irregularly alternating and that the



Standard 20 Words per Minute

Balanced 20 Words per Minute

Balanced 40 Words per Minute

Balanced 60 Words per Minute

Fig. 6

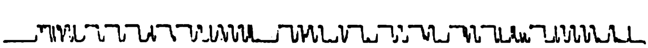
average values above and below the zero line are equal. It will also be observed that during the interval of a dot or dash that the signaling current is unidirectional over a longer period than is required for the residual currents produced by the atmospherics to reverse themselves. Hence the two currents are capable of further separation by means of the low pass filter *D* interposed between rectifiers and recorder.

The foregoing analysis describes in a general way what happens in this system when signals are being received alone and when static is being received alone. While a great deal more might be written about the action when both are received in combination since the particular phase relation between the signaling currents and the disturbing currents modifies the behavior of the system somewhat, yet it will be along the lines laid down and this part of

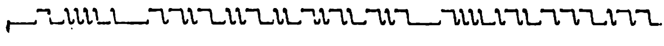
(* The particular initial phase relation between the heterodyne current and the individual natural waves.)

the operation can best be followed by an analysis of the tapes taken under conditions of practical working.

There are two bases upon which comparisons may be made between the standard method of signaling and the method herein described. One is to make a record of the balanced method at a speed at which the tape is just readable without error and to compare it with a record taken at the same speed with the standard method, estimating the extent to which the signal has been buried



Standard 40 Words per Minute



Balanced 100 Words per Minute

Fig. 7

in the second case. The other is to make a comparison of the relative speeds of the two methods at which it is possible to work without error.

The first method of comparison is perhaps the most spectacular and interesting to the engineer, particularly those who have spent much time on the problem. The second method is the one which

Standard 30 Words per Minute

Balanced 30 Words per Minute

Balanced 60 Words per Minute

Fig. 8

will appeal to the traffic manager and to all those who deal with the delicate balance between paid words handled, and overhead operating expense and cable competition.

The results obtained with either of these methods will vary depending on the type of atmospheric disturbances encountered. For example, assuming extremes, we might have a condition where the disturbances were due entirely to lightning strokes occurring in the immediate neighborhood of the station. Even though the balanced method were capable of reducing the effect

of these disturbances to a few per cent of their original value the overpowering effect of the residual would make its mark upon the tape and destroy the signal to the same extent as in the standard method, and no advantage would be gained. On the other hand the disturbances might consist wholly of grinders of about the same amplitude as the signal but present in such quantities as to bury it completely when using the standard method. These grinders, if reduced to a few per cent of their value, would disappear as a factor and a signal readable at any speed within the capabilities of the recorder would result. Between these two theoretical limits there lie the conditions of practical working, although both extremes are sometimes approached.

There are illustrated in Figs. 5, 6, 7, and 8, photographic reproductions of representative sections of records which have been taken over a long period of time under the varying conditions referred to and with both methods of comparison. The comparisons were made between the arrangement shown in Fig. 1 and the standard form of receiver now widely used in trans-oceanic work. This latter consisted of a tuned amplifier system for the 20,000-cycle current, comprising four tuned circuits arranged for a maximum of selectivity, a push pull detector system with separate heterodyne, and a low frequency amplifier whose output, when rectified by a simple valve, produced a current of 6–8 mil amperes in the coil of an R.C.A. standard siphon recorder. The comparisons were made on the non-directional antenna at Columbia University of a single wire about 100 feet high and 500 feet long. Signals were transmitted from a local oscillator feeding into the antenna and arranged to transmit either the normal type or the double frequency type of signal. The frequency used was about 20,000 cycles with a variation of 40 cycles. The ordinary Wheatstone automatic was used throughout. On account of the difficulty of operating two systems with two separate recorders simultaneously the records for comparison were made consecutively, usually within a few seconds of each other. In all practical cases this gives a sufficiently accurate comparison.

Fig. 5 shows some tapes taken according to the first method of comparison at three different speeds—20, 40, and 75 words per minute. In each case the record for the balanced method was taken first, the strength of the transmitted signal being adjusted to give just readable tape. As soon as this record was completed the record for the standard method was taken, both speed and signal strength remaining constant. These records were taken

under conditions encountered on the ordinary summer evening and speak for themselves.

Fig. 6 shows some records taken according to the second method of comparison. In each case the tape for the standard was taken first, the signal strength being adjusted to give a readable signal. The recorder was then connected to the balanced system, and, with the same signal strength, the speed of the

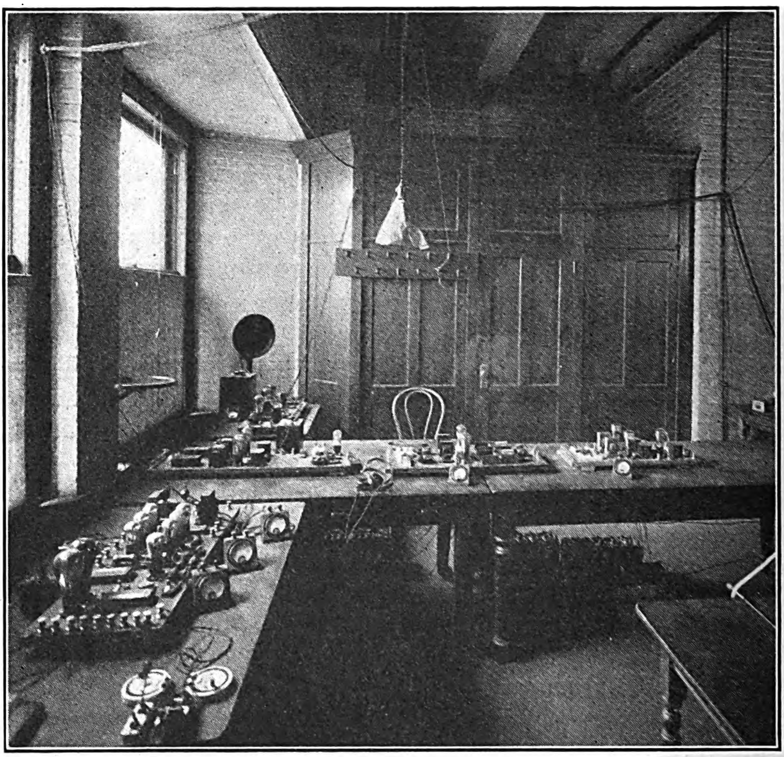


Fig. 9—Balanced Receiver Set-up in Hartley Research Laboratory.

Wheatstone increased until the limit of readability was reached. This set of tapes shows that a signal on the standard which is just readable at 20 words per minute is easily readable with the balanced method at 60 words per minute. Another set of tapes illustrated in Fig. 7 shows a signal readable at a speed of 40 words per minute on the standard reaching easily 100 words per minute with the balanced method. In this particular case the limit was not the atmospheric disturbances but the inability of the transmitter relays available to behave properly above this speed. The

operator who is the medium between a ragged tape and a typewritten copy. The criterion in facsimile transmission is one of intelligibility plus the question of how much permissible smudge may be forwarded to the customer on his message. In this last respect the difference between the two methods is even more pronounced than in the matter of speed.

In closing this paper I want to make an acknowledgment of the debt I owe to my old professor, Michael Idvorsky Pupin. Over thirteen years ago we began an investigation of the problem of atmospheric disturbances. For three years we continued it to the exclusion of all other work. The result was barren in the sense of arriving at a solution, yet the instruction that I received in electrical transients and the knowledge that I gained of that particular kind of transient which we designate as static, lies at the base of the present development. I want also to express my appreciation to Mr. Thomas J. Styles, whose assistance throughout the course of this work has been invaluable.

Discussion

Carl R. Englund: Some dozen years ago the subject of static was a very live and at the same time a rather mysterious subject. The mystery is in great measure gone now, and a realization of what can actually be done to mitigate static has tempered the interest which most of us feel in the subject while the fact that we still know little of the origin of static is of more theoretical than practical importance. Now, in the paper before us, we have a former line of static argument reopened and I am sure I am not alone in the feeling that this paper should be carefully scrutinized.

The earlier schemes for balancing static out while leaving the signal in, may perhaps be grouped in three classifications. Thus:

(a) Schemes where receptions from several antennas with unlike constants are directly combined to balance out the static.

(b) Schemes as above but combined for balancing after detection or rectification.

(c) Schemes combining the outputs of antennas either having different polar characteristics or spaced widely apart so as to obtain directive characteristics and balances.

We now know that schemes (a) do not work satisfactorily and that schemes (c) may be made to work satisfactorily. Schemes (b), however, in which category the Armstrong apparatus appears to be included, are more difficult to prove or disprove. For schemes (a) a balance is obtained only when both amplitudes and phases are correctly adjusted for all frequencies and one of the ideas justifying a study of schemes (b) has been that by rectification we in some measure remove the phase balance condition from the requirements for a static balance. As far as I know this scheme was first discussed by John Mills some eleven years ago and first tested by H. T. Friis about six years ago. Experiment showed that single impulses could be balanced out but overlapping impulses could not be balanced. Now for a pure impulse and two antenna systems, say, the transient currents flowing will be

$$\left.\begin{array}{l} I_1 = I_a l^{-\alpha_1 t} \sin \omega_1 t \\ I_2 = I_b l^{-\alpha_2 t} \sin \omega_2 t \end{array}\right\}$$

and if by adjustment of the circuit constants we make $\alpha_1 = \alpha_2$ and rectify, we shall have two rectified pulses having about the same shape or envelope and a balance adjustment can be made. But

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a second impulse occurring before the currents from the first impulse have subsided finds itself occurring in the two systems at a different phase with respect to the existing current, which adds in on the new transient, and the rectified pulses will no longer have a common shape. This simple line of argument is not rigorous when applied to static but gives a picture which can be used with discretion. Certainly the larger part of static fits more nearly the overlapping rather than the isolated impulse case.

Another method of attack on the static reduction problem is by the use of selective circuits. Here the mathematician can be of assistance and has already rendered aid. Koert's "Atmospbărische Störungen in der drahtlosen Nachrichtenübermittlung," 1924, Burch and Bloemsma *Phil. Mag.* **49**, 480, 1925 and Carson, *Bell Tech. Jl.* **4**, 265, 1925 all arrive at substantially the same result, by an application of the Fourier analysis with certain plausible assumptions. This result may be roughly stated as follows. The figure of merit of a selective system for a signal frequency band versus an arbitrary interference, is the ratio of the area under the resonance curve of the system occupied by the band to the total area under the resonance curve. The chief assumption required is that the frequency spectrum of the interference have no pronounced oscillations through the width of the resonance band, an assumption which is defensible.

The application of the foregoing to Armstrong's paper is evident. The fact that he uses a common antenna for two receptions does not seem to be vital, especially when used in connection with the balanced amplifier B and the circuit M, N, O, P, Q. Both of these latter are conjugate devices to split his double-frequency transmission into two quite separate channels so that the balancing after detection comes under schemes (b). At the same time that he utilizes this balancing scheme he has both ordinary resonant circuits and a low pass filter operating on the signals and the resulting selectivity should be marked and possibly nearly equal to the maximum useful selectivity his relatively low-frequency signal can use. Since schemes (b) have been found inoperative on practically all static in earlier work, and as both theory and practice demonstrate the pronounced reduction of interference possible by high selectivity on narrow band transmission of the type he used, it is fair to ask if his results are not in great measure due to selectivity alone. None of his signals occupied a frequency band much exceeding 100 cycles in width, and if I understand his paper correctly he has *not* compared balance and non-balance

conditions on signal transmission using the same total selectivity for both. I refer particularly to the low pass filter in his own apparatus. A mere comparison of the performance of a separate standard set with his apparatus is not a valid procedure for proving the operation of a static balancing apparatus.

Considering the importance of Mr. Armstrong's conclusions, the data given by him are rather too brief to be satisfactory. Record *F* does show a marked balance over records *D* and *E*, but if these are all to the same time scale as records *A*, *B*, and *C*, a low pass filter would remove practically all the static unaided in my opinion. Records *D*, *E*, and *F* are the only static-bearing records depending solely on a balance which he gives, and are not sufficient to permit an estimation of the part played by the "balance" factor in the improvement of the signal reception records 5, 6, 7, and 8. It is of course this "balance" in which we are above all things interested.

Mr. Armstrong in his second paragraph assigns the operation of the system to an energy balance valid for signals as against static. This can ordinarily be defended if qualified by the proviso that the energy is averaged over a sufficient interval, but experience shows that such intervals do not exist. Mr. Armstrong, however, rallies to his aid the idea that the static energies in two receivers operating at only slightly different frequencies will be so nearly identical in electrical form that a balance can be gotten. But the sharper the bridge balance necessary to resolve two adjacent frequencies becomes, the greater the ability of the system to pick out unbalance also becomes, and a net gain seems doubtful. It is surely to be hoped that Mr. Armstrong will give us further data where the separate roles of balance and selectivity are more clearly brought out. Especially to be desired are over all, or ether-to-siphon-recorder, selectivity curves for both sets.

AUTOMATIC VOLUME CONTROL FOR RADIO RECEIVING SETS*

By

Harold A. Wheeler

(Hazeltine Corporation)

Summary—*A receiving set is described in which the radio-frequency amplification is automatically controlled to give a nearly constant radio-frequency voltage at the detector, independent of differences in antenna signal voltage. This results in nearly uniform response at the loud-speaker from nearby and distant broadcasting stations and also reduces the effect of fading. The method employed consists in using the rectified carrier voltage to adjust the grid bias of the radio-frequency amplifier tubes. There are indicated the solutions of special problems that arise in carrying out this method.*

IN the present radio receiving sets employing high amplification, it is necessary to adjust carefully a "volume control" in order to reproduce signals of different intensities with the same audible intensity from the loud speaker. There are various devices which could be employed to regulate automatically the

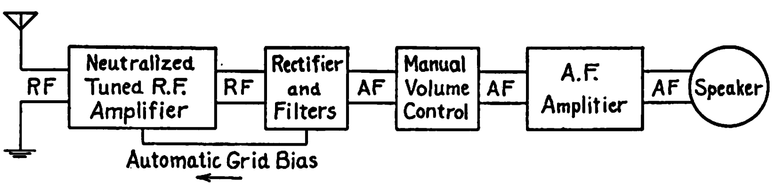

Fig. 1

amplification of the signal, some of which employ moving mechanical parts. It is the purpose of this paper to describe a simple electric circuit, without moving parts, in which the amplification is regulated automatically by the signal, and the loud speaker intensity reaches approximately the desired level for each signal, independent of the signal intensity and therefore irrespective of a reasonable amount of fading.

Any device to accomplish this object without introducing distortion of music or speech must operate by the signal carrier wave. Any variations in its intensity must be compensated by reciprocal variations in its amplification. The method to be described provides for controlling the radio-frequency amplifier, thereby maintaining the desired signal level in the detector or rectifier, audio-frequency amplifier and loud-speaker.

* Received by the Institute, October 6, 1927.
* Presented before the Institute of Radio Engineers, New York City, November 2, 1927.

Fig. 1 shows the outline of a set which has been constructed for broadcast reception, embodying this automatic volume control, comprising the following component sections. (1) A four-stage radio-frequency amplifier of the well-known Neutrodyne type, with UX 201-A tubes, the antenna circuit tuned by one dial and

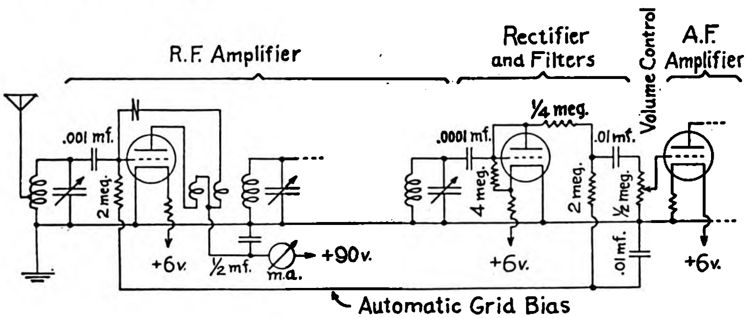

Fig. 2

the four coupling transformers tuned simultaneously by a second dial. The total amplification is controlled by varying the negative grid potential of the first three tubes. (2) A two-element rectifier with simple filter circuits to reject the radio-frequency currents

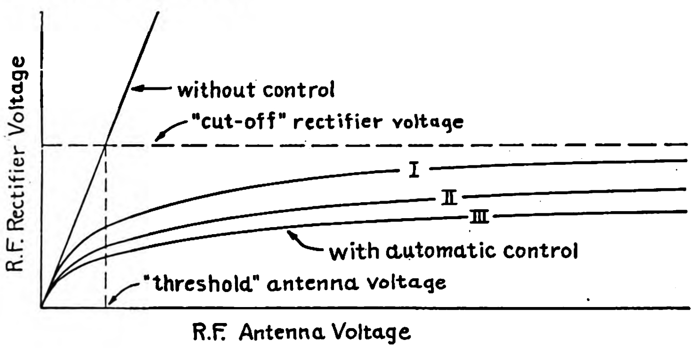

Fig. 3

and to segregate the direct and audio-frequency components of the pulsating rectified voltage. (3) A manual volume control in the form of a voltage attenuator connected to the grid of the first audio-frequency amplifier tube. (4) A four-stage audio-frequency amplifier and loud speaker. The entire set, excepting the last two audio-frequency stages, was enclosed in a grounded metal box divided into compartments, one for each tube with its preceding coupling circuit.

Fig. 2 shows the essential circuit details pertaining to the control system. The direct component of the rectified voltage, free of audio-frequency variations, is applied to the grids of the first three tubes. If the radio-frequency rectifier voltage could exceed a value of about ten volts, this automatic grid bias would thereby cut off the signal through the radio-frequency amplifier, so the rectifier voltage cannot exceed this value.

Fig. 3 shows graphically the comparison between the performance of the radio-frequency amplifier with and without the automatic control. With the system described, the rectifier voltage and audio-frequency voltages are nearly independent of the an-

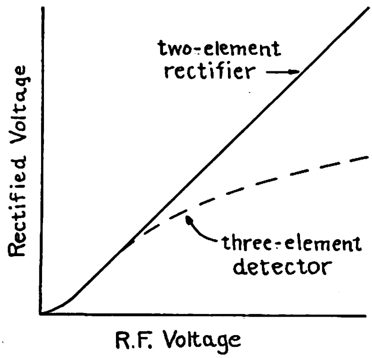

Fig. 4

tenna voltage, when the latter exceeds the threshold value. The curves I, II, and III show the performance of the system when the automatic grid bias is applied to one, two, or three tubes, respectively, of the radio-frequency amplifier.

The degree to which the signal can be cut off in one tube is limited by two factors. First, any error in neutralizing the grid-plate capacity permits signal current to pass through the tube, even when its mutual conductance is zero. Secondly, the sharp bend in the plate-current grid-voltage curve causes distortion of a strong signal on the grid, when the mutual conductance is reduced too far by the grid bias. In view of such limitations, it is undesirable to reduce the amplification ratio per stage below about 1/10 of its normal value. When controlling several tubes, these limitations become unimportant. The last radio-frequency stage is not controlled because it must supply as high as ten volts to the rectifier.

The properties of the two-element rectifier contribute largely to the simplicity of the control system. Fig. 4 shows the nearly

linear proportionality between alternating and rectified voltages in this form of rectifier, as contrasted with the irregular performance of the three-element detector. The signal modulation is rectified without distortion. Also the average rectified voltage is equal to the rectified carrier voltage, while with a "voltage-squared" detector the average rectified voltage is proportional to the average total power of carrier and sidebands. This last feature is worthy of mention in connection with the control system, since the automatic grid bias should depend only on the carrier amplitude, independent of the modulation.

With the circuit constants shown in Fig. 2, the time constant of the circuit which connects the rectifier to the grids of the control tubes is 1/40 second, so that the control system comes nearly to equilibrium in 1/20 second. This time can be reduced further if necessary, but is ultimately limited by the allowable reduction of the signal modulation at the lowest audio frequencies.

In consequence of the automatic control action, it becomes difficult to tune the receiving set accurately by ear to a desired signal. The amplification of the controlled tubes is decreased as the response to the signal is increased by tuning, and vice versa, so that the point of resonance is indicated by minimum plate current in the radio-frequency amplifier. Taking advantage of this fact, a milliammeter (*m.a.*, Fig. 2) is connected in the plate circuit of the first tube, to be used as a resonance indicator, and also to give an indication of relative signal intensities.

There is an incidental problem in supplying the plate current to all tubes of the set described from a common rectifier and filter system. In the controlled radio-frequency amplifier tubes, when operating at low plate current, the signal carrier is modulated appreciably by small fluctuations in the plate voltage. Such fluctuations are caused by the plate current pulsations in the audio-frequency amplifier. In the presence of a strong carrier wave, these two effects may cooperate to generate a low frequency oscillation. This disturbance may be avoided by reducing the internal output impedance of the rectifier-filter, by decreasing the amplification at low frequencies in the audio-frequency amplifier, or by using separate rectifier-filter systems to supply the plate currents of the radio- and audio-frequency amplifiers, respectively.

The performance of the automatic volume control as described can be summarized briefly as follows. A maximum variation of signal voltage in the ratio of 1:1000, corresponding to differences in distance, fading, or tuning, results in a maximum variation of

the rectified carrier voltage in the ratio of only about 1:3. This small variation, together with possible differences in the degree of modulation of different stations, can readily be compensated if necessary by adjusting the manual volume control for the audio-frequency amplifier, which also determines the "volume level" for the automatic volume control.

The name "Audiostat" has been selected for this device, by reason of its tendency to maintain the audible intensity at a constant value.

Attention might be called to British Patent 259,664 (Western Electric Co., July 14, 1925), in which a somewhat similar system is presented. This latter system is applied to a super-heterodyne receiving set, and is more involved in several respects than the system described in this paper.

It is desired to acknowledge the cooperation of the Howard Radio Company of Chicago, in whose laboratory the set described was assembled.

Discussion*

G. W. Pickard: Mr. Wheeler has confined his bibliography to a single reference, British Patent 259,664 of July 14, 1925. May I suggest that the Proceedings of the Institute of Radio Engineers is in this case a more fertile field than patent files, for in a paper entitled "Short Period Variations in Radio Reception," presented before this Institute on December 12, 1923, and published in the Proceedings of April, 1924, will be found a description of the exact system of Mr. Wheeler's paper.

For convenience, I will give the quotation from my 1923 paper:

So far as the varying intensity of the sound in the telephone receiver is concerned, there are several simple expedients which will markedly smooth out reception from a distant station. Thus, the grids in a radio-frequency amplifier train may be floated on a fairly large condenser, shunted by a high resistance. When reception is weak, the grids assume a small negative potential, and amplification is at a maximum. When the input rises, a large negative charge is built up on the grids, and amplification is reduced. *Similarly, a separate rectifier connected to the output end of the radio-frequency amplifier and to the grids will have the same effect.* (Italics mine. G. W. P.)

By accident rather than by design, some of the resistance-coupled radio-frequency amplifiers of 1917–1918 achieved a certain degree of automatic volume control. In these amplifiers rectification went hand-in-hand with amplification, with the result that a strong signal would build up so large a charge on the grids that amplification was greatly reduced.

Mr. Wheeler mentions the difficulty in tuning such a receiver by ear. This difficulty is due to his choice of a too-small time-constant for the rectifier-grid circuit. Inasmuch as the more important fading periods of broadcast reception are of the order of a minute or more rather than seconds or fractions of a second, it is not necessary or even advisable that the control should operate within a small fraction of a second. In my work with this form of receiver, I made the time-constant of the control about ten seconds, which gave ample time for tuning.

It is not likely that any system of automatic volume control will make a distant station behave just like a local. When one is located in the zone of most violent fading, that is, at such a distance from the transmitter that the direct and refracted waves are of the same amplitude, the field intensity at the bottom of a deep fade is usually below the disturbance level. Under these conditions, I have found that a receiver with automatic volume

* Received by the Institute, November 2, 1927.

control may maintain an approximately constant signal output from the loud speaker, but this will be periodically obliterated by the rising and falling noise level.

E. Bruce†: Several years ago during the ship-to-shore experiments conducted by the Western Electric Company, Mr. H. T. Friis, now of the Bell Telephone Laboratories, employed the automatic electrical gain control disclosed in British patent No. 259,664 to which Mr. Wheeler's paper refers. In this connection, a circuit suggested by Mr. Affel of the American Telephone and Telegraph Company disclosed in U. S. patent No. 1,511,015, October 7, 1924 is of interest.

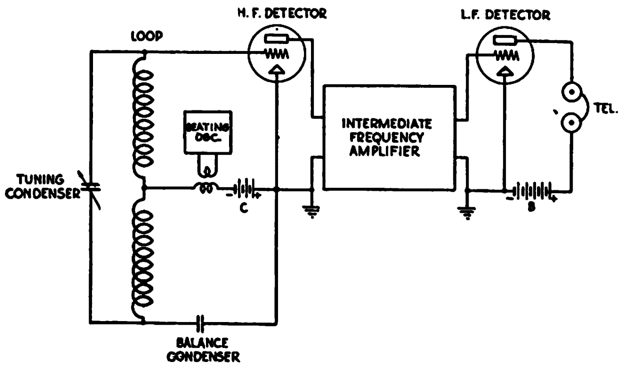

Fig. 1

The writer has devised a method of automatic gain control which is extreme in its constructional simplicity and at the same time avoids the range limitations experienced by Mr. Wheeler and also mentioned by Mr. Friis. Since this material has never been published, it seems appropriate, at this time, to present it to the Institute for consideration along with the method described by Mr. Wheeler.

The basic idea involved is to cause the rectified output to operate on the frequency changing device of a double detecton or "superheterodyne" system. Since the output of this device occurs at the intermediate-frequency, no signal output is possible when this device is made to be inoperative. In other words, we obtain a practically infinite operating range. This is in marked contrast to operating on amplifier tubes which require elaborate and accurate balancing schemes to prevent the signal from passing

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† Bell Telephone Laboratories, New York City.

via the inter-electrode capacities of the tube. Mr. Wheeler has pointed out that the range of such signal reduction is limited to the practical accuracies to which these balances may be adjusted.

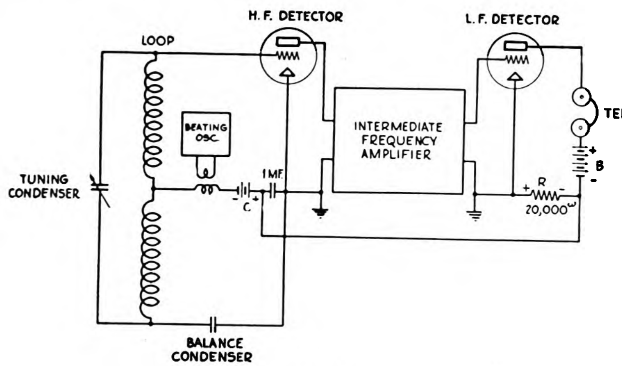

Fig. 2

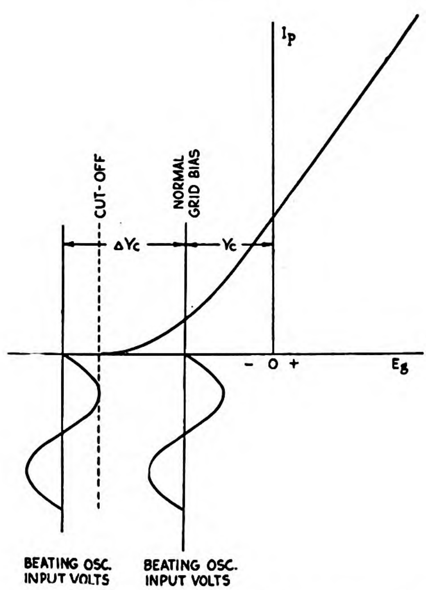

Fig. 3. Characteristic of High-Frequency Detector.

Fig. 1 shows an approved form of double-detection receiver. Fig. 2 indicates how it may be provided with an automatic control by simply adding a one micro-farad capacity and a 20,000-ohm resistance.

Fig. 2 shows that the rectified output of the low-frequency detector causes an increase in the IR drop across the 20,000-ohm resistance. This IR drop furnishes an additional negative bias to the high-frequency detector, driving that tube toward plate current cut-off. At cut-off the amplification of the receiver is totally destroyed.

Fig. 3 shows the change in negative bias necessary to reach cut-off. Referring to the figure,

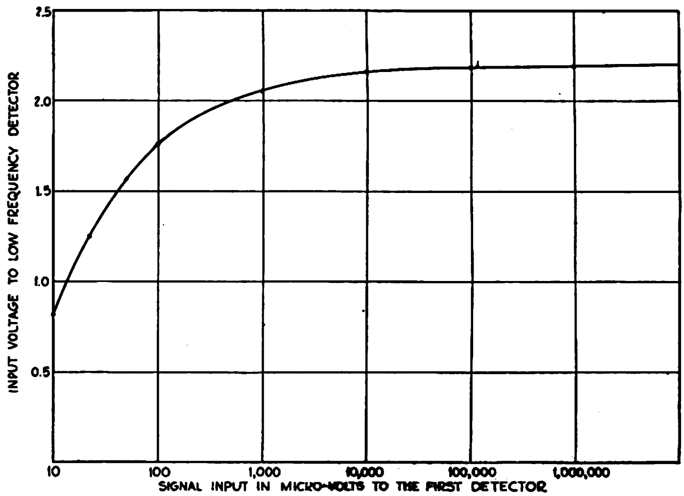

Fig. 4

$$\text{lim. } \Delta V_c = V_s + V_{bo} + \frac{V_b}{\mu} - V_c$$

where ΔV_c = change in grid bias.

V_s = peak signal voltage at detector input.

V_{bo} = peak voltage of beating oscillator input.

V_b = plate battery voltage.

μ = voltage amplification of tube.

V_c = normal negative grid bias.

Let us assign practical values to these terms respectively in the above order, then

$$\text{lim. } \Delta V_c = \text{ negligible } + 1 + \frac{45}{6} - 6 = 2.5 \text{ volts.}$$

ΔV_c is the ultimate limit (signal voltage small compared with beating oscillator) of the biasing voltage that need be provided by the drop across the resistance R.

$$\lim.\ \Delta V_c = I_r R.$$

$$I_r = \frac{\lim.\ \Delta V_c}{R} \times 10^6 \text{ microamperes} = \frac{2.5 \times 10^6}{2 \times 10^4} = 125 \text{ microamperes.}$$

An average N tube voltmeter characteristic shows that a change in plate current of 125 microamperes represents an input of 2.5 volts for a plate circuit load of 20,000 ohms. It is therefore concluded that the input to the low-frequency detector can never exceed 2.5 volts for any practical signal.

Fig. 4 is a measured characteristic of the signal input in micro-volts to the first detector vs. input voltage to the low-frequency detector. The intermediate frequency amplifier possessed a voltage gain of 120,000. If we start at a level of 1000 micro-volts and increase this signal 1000 times, the output will only increase by about 6 percent.

With the arrangement as described, the desired output may be manually adjusted by altering the value of the negative biasing battery marked C in Fig. 2.

An automatic gain control of the kind described has been found to operate very satisfactorily. In the circuit used, the flat characteristic has been improved through the use of detector tubes having a voltage amplification constant of 30 instead of 6.

In the present state of development of the automatic gain control, there are two serious limitations: First—As the gain of the set rises and falls, in order to compensate for a fading signal, it is accompanied with a corresponding rise and fall of static and inherent set noise. This is quite disturbing to intelligibility. Second—It is common experience that speech quality is poor during the minimum of a fading period. An automatic gain control is incapable of remedying this situation.

RADIO COMMUNICATION*

By

Senatore Guglielmo Marconi, G.C.V.O., D.Sc., LL.D.

I very much appreciate the honour of being able to read this address before members of the American Institute of Electrical Engineers and of The Institute of Radio Engineers, especially as I know that in America radio science is more deeply investigated, more universally understood, and more generally utilised than in any other country on earth.

I also cannot but cherish always the recollection that the American Institute of Electrical Engineers was the only technical and scientific body which more than twenty-five years ago first believed in me and endorsed my statement that in December, 1901, I had succeeded in getting the first radio signals across the Atlantic Ocean, the first distinguished and authoritative Society enthusiastically to celebrate the event and to extend to me their generous support and valuable encouragement.

It is a further satisfaction to me to realise that as a result of recent discoveries and inventions the subject of radio communication is today attracting more world-wide interest and attention than any other advance of physical science and of electrical engineering.

In the early days of "Wireless," when electromagnetic waves were first beginning to be employed for practical purposes, we spoke only of "wireless telegraphy," or "radio telegraphy," but with the advance of the art these waves have been more and more widely utilised not only for telegraphy but also for telephony and broadcasting, direction-finding at sea and in the air, for the control of mechanisms and the ignition of explosives at a distance, principally for war purposes and, more recently, also for the transmission of line drawings, photographs and facsimiles, and finally, for television, which is now, I believe, finally emerging from the laboratory stage.

I hope I shall not be thought too visionary if I say that it may perhaps be possible that some day electromagnetic waves may

* Delivered before the American Institute of Electrical Engineers and The Institute of Radio Engineers, New York City, October 17, 1927.

also be used for the transmission of power, should we succeed in perfecting devices for projecting the radiation in parallel beams in such a manner as to minimize their dispersion and diffusion into space.

The achievements and possibilities of radio have already become so vast, so far-reaching, and their theory so complex and undergoing at the present time such a bewildering process of evolution that you will easily understand that, did I not confine myself to the generalities of even a small part of my subject, I would find it quite impossible to keep the length of my address within practical limits.

It would also be quite useless for me to endeavour to describe at any length the general achievements and utility of radio in a country where so very much is already known of this art and science and where such gigantic strides are being made in its practical application and scientific development.

I shall, therefore, necessarily be unable to dwell upon the valuable research work on the now all-important subject of short waves which has been carried out here, particularly by the engineers of the Radio Corporation of America, but this has already, in part, been described in an admirable paper by Messrs. H. E. Hallborg, L. A. Briggs, and C. W. Hansell, which was published in the June number of the PROCEEDINGS of The Institute of Radio Engineers.

I, therefore, propose to limit myself to referring briefly to the development and utilisation of this latest and most important evolution of radio science, which has already had the effect of compelling us to modify radically our views in regard to the practice and theory of long-distance transmission.

For this purpose I shall confine myself principally to a brief historical sketch of the investigations carried out by myself and my assistants on the subject of short waves and to describing some of the strides that have already been made in their application to radio communications over long distances.

I feel that I must here repeat my belief that we are as yet far from being able to assert that radio is based on well-understood foundations, unless, of course, we should go back to the long-distance technology of the past which, to my mind, has become more or less obsolete when applied to present-day long-distance practice.

Whatever degree of perfection may have already been achieved

in the design and construction of "wireless" stations and radio apparatus, we must realise that we still know too little of the true mechanism governing the propagation of the waves, and of the properties and behavior of the space which they traverse.

Speaking generally, it seems to me that latest developments tend to show us that four or five years ago radio engineers thought they knew much more about the subject than, perhaps, we think we know today. Laws and formulas were announced and accepted showing which wavelengths were best adapted for various distances for both day and night transmission and indicating what amounts of power would be necessary in order to enable us to communicate with a fair degree of regularity over any given distance. But, unfortunately, the logical application of these laws and formulas brought us to the necessity of employing for long-distance work such enormous and expensive antenna-systems and such large amounts of power as to make radio transmission so costly in capital expenditure and operation that it would hardly compete economically with modern cable and land lines.

The study of what are now termed short electrical waves can be said to date from the time of the discovery of electromagnetic waves themselves, that is, from the time of the classical experiments of Hertz and his contemporaries nearly forty years ago: for Hertz used short waves in his laboratory when he first conclusively proved that electrical waves existed, and that they were subject to the same laws as the waves of light in regard to reflection, refraction, diffraction, interference, and speed of propagation.

I might also, perhaps, recall the fact that in my own earliest experiments thirty-one years ago I was able to demonstrate the transmission and reception of intelligible signals through space over a distance of $1\frac{3}{4}$ miles by means of a directive system employing waves of only about 1 meter in length, whereas at that time, by means of the antenna or elevated wire system employing much longer waves, I could only, curiously enough, get signals over a distance of about one mile and a half.[1]

The progress which has, however, been made subsequently with the long-wave system was so rapid and so spectacular in regard to distance, and the results available so easily applicable to the urgent needs of shipping, that it diverted all research from short waves, especially as it appeared, as indeed was proved,

[1] See paper read before the Institute of Electrical Engineers in London, March, 2, 1899.

that by efficiently utilising waves longer, and longer than those of about 150 meters—which were the first to be employed for any considerable distance—the ranges over which it was possible to communicate were steadily increased and the absorption caused by the effect of sunlight decreased and later, by the use of the longest waves, finally overcome.

This neglect of short waves was, I think, regrettable, for, notwithstanding the intense radio research that has been carried out in most countries for the last twenty-five years at least, it has been left to us only recently to discover that these waves possess most valuable and unsuspected qualities in regard to world-wide transmission and that they are capable of results unobtainable by the lower frequency system which, up to almost the present day, has held the field for all long-distance radio communication.

Since my early experiments carried out in 1896–97 and for a very long period of years afterward, no serious research work was carried out, or at least published, so far as I can ascertain, in regard to the application of very short waves to radio purposes.

Research along such lines did not appear promising: short waves were not easy to produce or to detect with the means then at our disposal, and up to recent times the power that could be put into them was small. This, together with the erroneous but general belief of the high attenuation of the waves over even short distances, deterred experimenters from entering this new field of research.

Some years ago, during the World War, I could not help feeling that we had perhaps gotten into a rut by confining all our researches and all our tests to long waves; that is, to waves of hundreds of thousands of meters in length; especially as I realised that, in accordance with theory, it would be practically possible only by the use of short waves to project the radiation in narrow beams in any desired direction instead of allowing it, as had always been done, to spread and dissipate in every direction.

I was greatly impressed by the advantages that such a system might possess for point-to-point communication, by the possibility which it would afford of reducing tapping and interference even if several stations were worked in the same area, and also by the possibility of a better and more logical utilization of the energy radiated from the transmitter.

My doubts were as to whether atmospheric absorption, the interposition of obstacles, and the curvatures of the earth would

not result in always limiting the distance of useful operation to a few score of miles, but I hoped that through the concentration of energy brought about by the utilization of efficient reflectors and, perhaps, by some unknown yet beneficent effect of the upper conducting layer, it might, nevertheless, be possible to effect communication across not inconsiderable distances.

This line of research was taken up by me in Italy early in 1916, and in subsequent development work during that year and afterwards I was most valuably assisted by Mr. C. S. Franklin, of the Marconi Company.

Mr. Franklin, under my direction, followed up the subject with great thoroughness, and the results of several years of our investigations were described by him in a paper read before The Institution of Electrical Engineers in London on the 3rd of May,

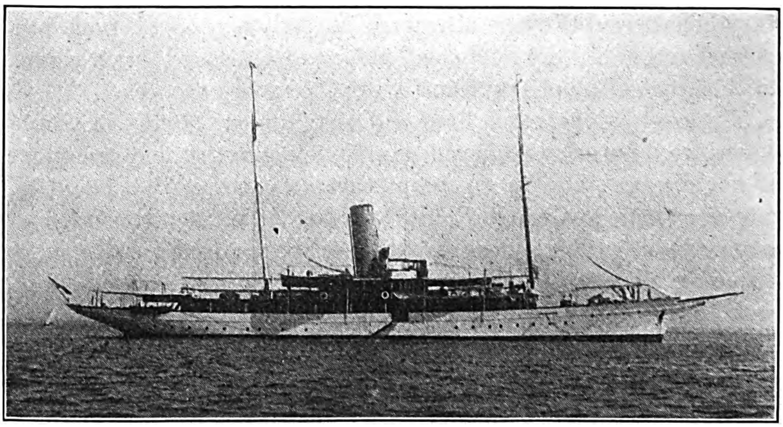

Fig. 1—Senatore Marconi's Yacht *Elettra*

1922, and also by me in an address delivered before a joint meeting of the American Institute of Electrical Engineers and The Institute of Radio Engineers in New York on the 20th of June of that same year.

The results obtained up to that time definitely convinced us of the enormous advantage to be gained by the use of suitable reflectors at both the transmitting and receiving stations. The tests were carried out with very small power and with waves from between 2 and 15 meters in length, up to distances of about 100 miles; but I should point out that at that time there was nothing to indicate to us that these distances constituted the limit range of

the waves thus employed. We did, however, ascertain by a number of careful measurements that the energy received when suitable reflectors were used at both the transmitting and receiving ends could be 200 times that of the energy received when no reflectors were employed.

Systematic tests, the object of which was to ascertain the range and capabilities of short waves over varying distances, were commenced by myself and Mr. C. S. Franklin in the spring of 1923 between a small experimental transmitting station situated at Poldhu in Cornwall and a special receiving station installed on the Steam Yacht *Elettra.*

The results obtained from these tests went far to convince me that short electric waves possessed qualities which, up to that time, had remained unknown and that this new line of investigation was opening up a vast field of profitable research full of undreamed-of possibilities.

The principal objectives aimed at in the experiments carried out between Poldhu and the S. Y. *Elettra* were:

(1) To ascertain the day and night ranges and reliability of signals transmitted on wavelengths of less than 100 meters, possibly over considerable distances, with or without the use of reflectors or directional devices.

(2) To investigate the conditions which might adversely affect the propagation of short waves, such as the interposition of land or mountains between the two stations, and also how the night or day ranges varied with the wavelength employed and the power utilized.

(3) To investigate and determine, if possible, the angle and spread of the beam of radiation emitted when employing a transmitting reflector, especially with a view to the possibility of establishing long-distance directional services.

A wavelength of approximately 97 meters was first employed, with a power of 12 kilowatts in the aerial, and during our journey, in the course of which ports and places in Spain, Morocco, Madeira, and Cape Verde were touched, it was ascertained that, with the power and wavelength employed, signals could be reliably received during daylight up to distances of 1,250 nautical miles.

In carrying out these tests I first noticed that it is by no means correct, in dealing with waves of approximately 100 meters, to refer to distances covered during daylight as "day ranges," because the strength of the signals which could be received varied definitely

and regularly in accordance with the mean altitude of the sun over the space or region intervening between the two stations.

The night signals came in always with great strength and remarkable regularity up to the maximum distance to which the yacht was able to proceed on that occasion, which was as far as the Cape Verde Islands, situated at 2,320 nautical miles from Poldhu. The strength of the signals received here at night left no doubt in my mind that their practical range was very greatly in excess of that distance.

I believe that I am right in saying that up to that time the general impression prevailing among most technical experts in regard to the behavior of very short waves was:

(1) That their range during day-time would be very short.

(2) That their night ranges, although occasionally considerable, would be exceedingly variable and freaky and subject to long periods of "fading," rendering their use altogether too unreliable for practical purposes in long-distance commercial working.

(3) That any considerable stretches of intervening land, especially if mountainous, would greatly reduce the distance over which it might be possible to communicate.

Our 1923 tests brought me absolutely and definitely to the conclusion that such opinions and beliefs were wrong in so far as they might concern the behavior of waves of about 100 meters in length, for we discovered:

(1) That the daylight ranges were not by any means inconsiderable and, in fact, proved to be much greater than had been anticipated.

(2) That the night working was very much more reliable than had been believed possible; that "fading" was not at all so serious as had been anticipated and that the great strength of signals received indicated that the night range would probably be much greater than anyone, myself included, had ever before expected or anticipated.

Moreover, a fact of great practical value was also observed, and this was that even in the tropical countries the atmospheric electrical disturbances, termed "static" or "X's," were invariably much less troublesome and severe when receiving on short waves than those experienced with the much longer waves which, up to that time, were being exclusively used for all long-distance work.

The results of these tests were set forth in a technical report drawn up at the time, and were also described in detail and pub-

lished in a paper which I read before the Royal Society of Arts in London on the 2nd of July, 1924. In that paper I ventured to predict that it would be possible by means of short wave directive stations of small power to send and receive a far greater number of words per 24 hours over world-wide distances than would be practicable by means of the existing or the then proposed powerful and expensive long-wave stations. This, at the time, may have seemed a bold statement, but I felt sure it was going to be justified by results.

Further tests and experiments were carried out in February and March, 1924, with the object of determining the maximum practical ranges of these waves, and we found that, while the day range of a 92-meter wave was about 1,400 miles, *i. e.*, greater than the day range of a 97-meter wave, strong and fairly reliable signals could be received during the dark or semidark hours not only in the United States but also in Australia; that is, over world-wide distances.

The results were so encouraging that I was tempted shortly afterwards to try a telephony test to Australia, which was quite successful (30th of May, 1924).

In August and September of the same year another series of tests was carried out between the Poldhu station and the S. Y. *Elettra*, with the object of studying the behavior of still shorter waves over long distances, in order to ascertain whether it might not be possible to overcome in some measure at least the curtailment of the hours of working brought about by the effects of daylight, for, of course, we realised that this limitation of the period of operation to practically only the hours of darkness constituted the principal drawback to the possible general adoption of these waves for commercial and practical purposes.

Experiments were, therefore, conducted over varying distances with four wavelengths of 92, 60, 47, and 32 meters respectively.

These tests enabled us to discover the important fact that for long distances the daylight range steadily increased as the wavelength was reduced below 92 meters, the 32-meter wave being received with ease all day at Beyruth, Syria, over a distance of 2,100 miles, while the 92-meter wave began to fail over this track during daylight at a distance not much in excess of 1,000 miles.

During these tests the 60-meter wave appeared to be slightly better than the 92-meter wave during daylight, the 47-meter wave still better, and the 32-meter wave very much better.

From the result of these experiments we naturally presumed, and later experience confirmed our anticipation, that still shorter wavelengths would show still greater daylight range and further tests carried out by ourselves and other workers not only proved this to be a fact but also showed that very short waves, while being capable of working over the greatest distances during daylight, had but a comparatively short and unreliable range of action during darkness.

This discovery, which has brought about a reversal of what was noticed in regard to wavelengths longer than 200 meters, apart altogether from its enormous practical importance, gives rise to scientific questions of the highest interest and importance and requiring theoretical explanation as to how these waves can travel, even right around the world.

I do not intend, on this occasion at least, to indulge in any theoretical hypothesis or theory, as I much prefer to confine myself to the description of observations and of what I believe to be facts, leaving to others to arrive at the most valuable theoretical inferences which may be deduced from them.

In October, 1924, transmission tests were carried out on a wavelength of 32 meters from England to specially installed receivers at Montreal, Long Island (New York), Buenos Aires, and Sydney (Australia), and it was at once found possible to transmit messages when utilising only 12 kilowatts or less at the transmitter to all these distant places even when the whole of the great circle track between them and England was exposed to daylight.

With Australia, however, it is only possible to have a track from England completely exposed to daylight for 2 or 3 hours at a time, and, furthermore, the scientific aspect of the test is complicated by the fact that the waves may have several ways of getting round the earth with comparative ease, as a large part of Australia is not very far from the antipodes of England.

The Australian tests showed, however, that it was possible to get through for about 23½ hours out of the 24.

Numerous other tests were carried out with various far-distant countries, including Japan, from a small power station at Chelmsford, England, which utilised only about *one-fifth of a kilowatt* of antenna energy, the object being to test still shorter waves. It was thus noticed that a wavelength of 10 meters was about the shortest which enabled signals to be detected at Sydney in Australia, and then only during the time when practically the whole of

the great circle track between England and Australia was exposed to daylight.

I should point out that these particular tests were carried out without reflectors at either end, the sole object being to ascertain the range and general behaviour of these waves over long distances.

In the directive experiments I carried out in Italy and in England, in 1916 and subsequent years, the reflectors consisted of a number of vertical wires of suitable length parallel to the antenna and spaced around it on a parabolic curve, of which the transmitting or receiving antenna constituted the focal line. The aperture

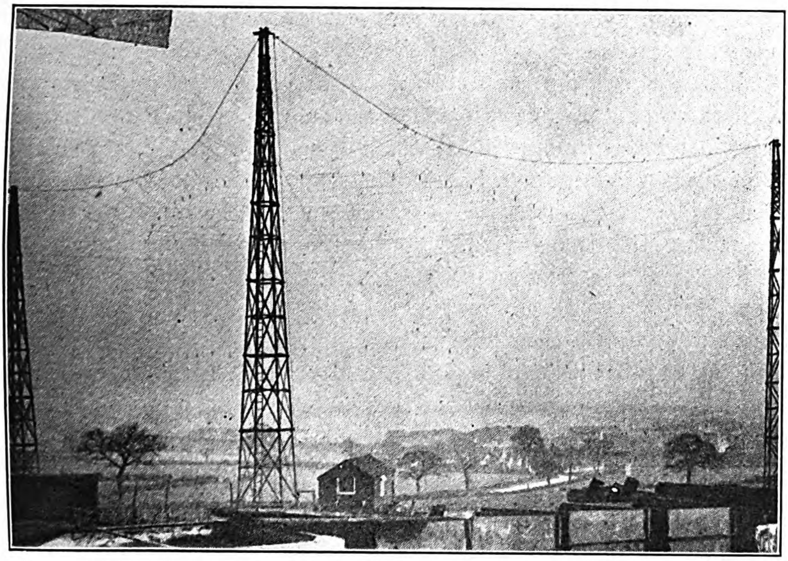

Fig. 2—Parabolic Reflector: Hendon

of the reflector was always made to be not less in width than 2 or 3 wavelengths.

Suggestions for utilising reflectors of this kind were made by Brown in 1901 and by DeForest in 1902, but many essential conditions necessary for efficiency were apparently not realised at that time, which fact may explain why no practical application of their arrangements was made.

Since 1916 various patents have been taken out by myself, Mr. C. S. Franklin, and other workers for reflectors and directive antennas, but in the commercial shortwave stations which have been constructed by the Marconi Company for the British and

other governments, and which are now in operation, an arrangement patented by Mr. C. S. Franklin is employed.

In this arrangement the antennas and the reflector wires are disposed so as to constitute grids parallel to each other, the aerials being energised simultaneously from the transmitter at a number of feeding points through a so-called "feeder system," and in such a manner as to meet the condition that the phase of the oscillations in all the wires is exactly the same.

It has been proved by calculations confirmed by experiments that the directional effect of such an arrangement is a function of its dimensions relative to the wavelength employed.

A similar system of aerials and reflecting wires is used at the receiving stations.

For a more complete account of our investigations, together with a more detailed description of the general principles on which my short-wave directional system is based, and also in regard to the apparatus employed, I would refer you to my papers read before the Royal Society of Arts on the 2nd of July and the 11th of December, 1924, to the paper read before the American Institute of Electrical Engineers and The Institute of Radio Engineers in New York on the 20th of June, 1922, to my "James Forrest" lecture delivered before the Institution of Civil Engineers in London on the 20th of October, 1926, to my paper "Le Radiocomunicazioni a Fascio" published in the Nuova Antologia of Rome in its issue of the 16th of November, 1926, to Mr. C. S. Franklin's paper read before the Institution of Electrical Engineers in London on the 3rd of May, 1922, and also to an article by Dr. J. A. Fleming which appeared in the *Wireless Engineer* of London in July of this year.

As I have already said, it has been proved long ago by calculations confirmed by observations that the directional effect of a radio reflector is a function of its dimensions in relation to the wavelength employed. Hence it follows that the dimensions of the reflector can be reduced proportionately to that of the wavelength and, therefore, that the cost is much lower and the space occupied much smaller for wavelengths of, say, 20 meters than for wavelengths of 90 or more meters.

The same calculations show us that the dimensions of a reflector for really long waves would be so enormous as to render its construction impracticable and impossible both for economic and engineering considerations.

Early in 1924 the British Government began to consider seriously proposals made by the English Marconi Company for the employment of the short wave directive system, now generally known as the "beam system," in order to satisfy the long-expressed desire of the dominions for a rapid and efficient means of radio communication with the mother country, and in July of that year a comprehensive agreement was entered into between the British Post Office and the Company for the construction of radio stations on the beam system in England and in the British Dominions, to

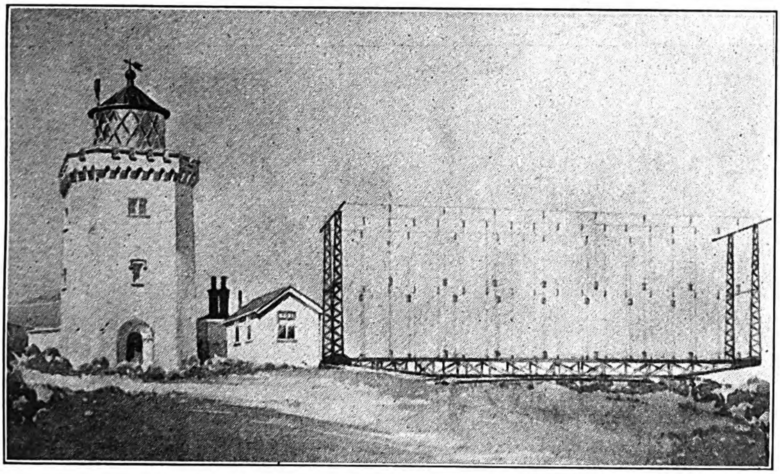

Fig. 3—South Foreland Revolving Beam

be operated in England by the Government, which would be capable of ensuring a high-speed commercial service from and between England and Canada, South Africa, India, and Australia.

This contract stipulated that the transmitting stations should dispose of a power of 20 kilowatts to be delivered to the anodes of the tubes used for generating the oscillations, that the aerial and reflector system was to be designed so as to concentrate the emitted waves within an angle of 15 degrees on either side of the axis of transmission, the energy emitted beyond this angle not to exceed 5 per cent of that on the axis, the receiving stations to have a similar aerial and reflector system designed to have its maximum receptivity in the direction of the corresponding station.

The conditions in regard to the speed of working were exceptionally stringent and severe.

The stations for corresponding with Canada were to be capable

of accurately sending and receiving at the same time to and from Canada at the rate of one hundred five-letter words per minute (exclusive of any repetitions necessary to secure accuracy), during a daily average of eighteen hours.

The stations for communicating with South Africa were to be capable of maintaining the same daily rate of speed and accuracy for eleven hours.

The stations working with India had to meet the same requirement for twelve hours and those working to and from Australia for seven hours daily.

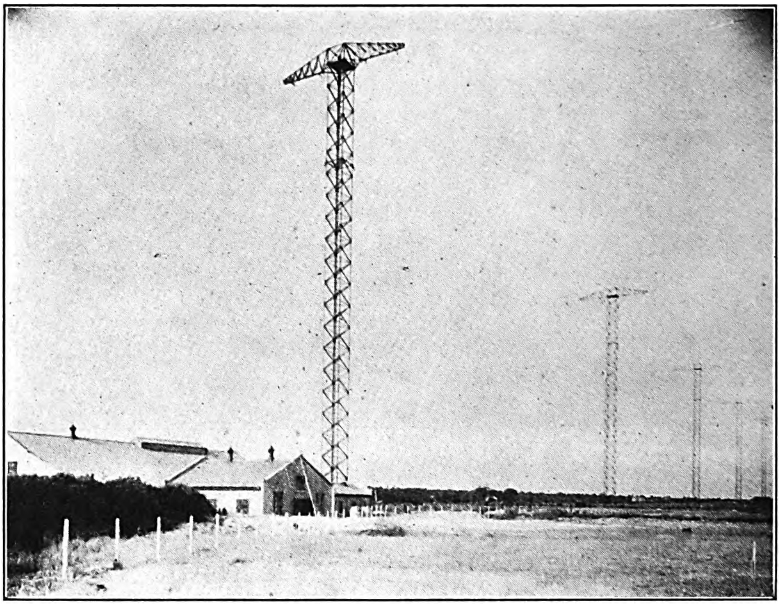

Fig. 4—Bodmin Beam Station. South African Masts

The Company was to give the Post Office a practical demonstration of actual working for seven consecutive days, which would prove to the satisfaction of the Government engineers that these severe conditions could be fulfilled.

In November of last year the Canadian circuit passed its official test, and in March, May, and August of this year the Australian, the South African, and the Indian Stations also completed their official communication tests and are now carrying on an important commercial public service which has already had the beneficial

effect of bringing about a reduction of the telegraphic rates between England and her most important dominions.

It has been found that, in regard to the stations communicating with Australia, South Africa, and India, a daily average of over twenty hours of high-speed communication is attainable, and that the spread of the beam is much narrower than had been specified in the Government requirements.

These results, which are truly unprecedented in the history of

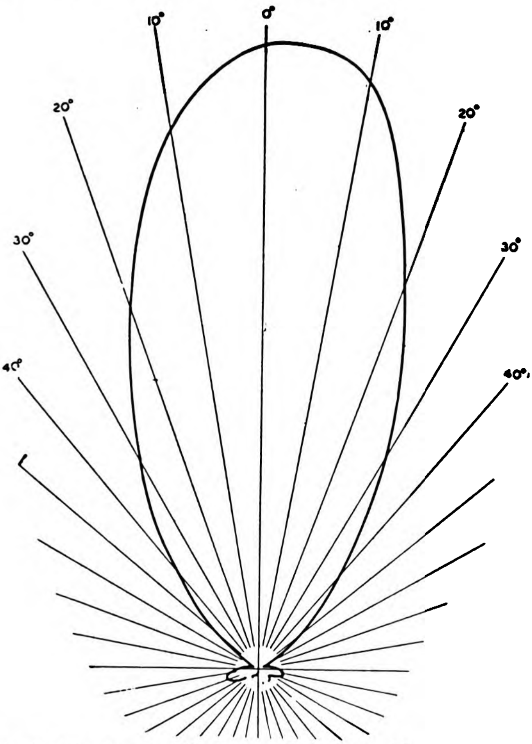

Fig. 5—Polar Curve, Hendon. Reflector, 28 Meter Aperture, 14.8 meter wave. Measured on circle of 31 meter radius.

radio communication, have been of the greatest possible satisfaction to myself and my co-workers, because they go to prove that our faith in the new system, which for years we upheld in the face of much scepticism and criticism, was not altogether misplaced.

The British Government has already expressed through its executive officials its high appreciation of the success of the new radio system.

Colonel T. F. Purves, Engineer-Chief of the British Post Office, in a letter of the 26th of August last addressed to the Marconi Company stated:

"All four stations erected by your Company providing direct telegraph communication with Canada, South Africa, India, and Australia have now been completed. It is with pleasure that I take this opportunity of offering my congratulations on the high degree of technical skill and resource displayed in successfully surmounting the many novel difficulties encountered in the work of carrying to fruition this great new development of the radio art."

The Rt. Hon. W. Mitchell-Thomson, British Post-Master General, also wrote on the 5th of October as follows:

"The introduction of the beam system will have far reaching effects in reducing the cost of long-distance communications by wireless telegraphy and I cordially congratulate your Company on a notable achievement."

The English stations for working with Canada and South Africa are situated at Bodmin, and at Grimsby for working with Australia and India; the receiving stations being respecitvely at Bridgewater and Skegness.

The corresponding stations in Canada are at Drummondsville and Yamachiche not far from Montreal; in Australia at Bellan and Rockgank, near Melbourne; in South Africa at Klikheuvel and Milnerton, near Cape Town, and in India at Kirkee and Dhond, near Poona.

The stations are all worked by what is termed "remote control" through land lines from the Central Telegraph Office in London, and the received signals are strong enough to work high-speed recording and even printing instruments also in the same office in London.

In India and in the other dominions the same system of direct control is in operation between the Telegraph Offices in Montreal, Melbourne, Cape Town, and Bombay with the corresponding radio sending and receiving stations.

The aerial and reflector system at each station is supported by a row of masts so arranged that the great circle bearing to the distant station with which each particular aerial system is intended to communicate is at right angles to the line of masts. The design of the aerial and reflector system is substantially the same at each station. The masts are spread at a distance of 650 feet from each other.

In the case of the Canadian, South-African, and Indian services, each reflector and aerial system is suspended on a line of five lattice steel masts, each 287 feet high, but in the case of the Australian stations where a single wave is being used there are only three masts and their height is 260 feet.

The Australian system has the further difference that it is constructed with an aerial system on either side of a central reflector so that it is capable of transmitting to Australia in the direc-

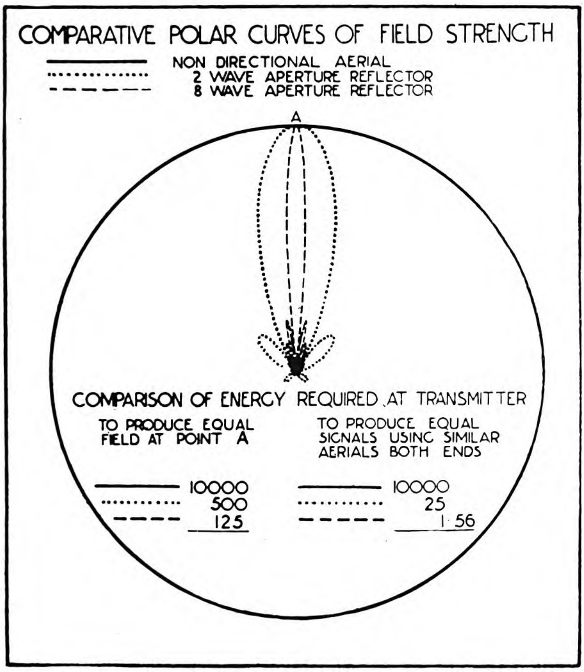

Fig. 6—Comparative Polar Curves on Beam and Round Transmitter

tion of either the easterly or westerly great circle. This arrangement has been made as a result of the experience which we gained when carrying out our preliminary short-wave tests with Australia at the beginning of 1924.

It was then found that the position and altitude of the sun had an effect as to which route the waves preferred to follow, and that during the morning period in England the waves preferred to travel from England to Australia in a westerly direction across the Atlantic Ocean following the great circle along the longest route which is approximately 14,000 miles; whereas, during the

afternoon and part of the night period the waves travel best in an easterly direction over Europe and Asia following the shortest great circle route which is about 10,000 miles. It is the practical application of this discovery which has resulted in the construction of the Australian transmitters and receivers so that transmission or reception can take place in either direction as required.

In all the sending stations the radio-frequency current is conveyed to and from the aerials through what I have already referred to as a "feeder system" consisting of concentric copper tubes air-insulated from each other to reduce loss. The outer tubes are earthed and carried on iron standards driven into the ground. The inner tubes, which constitute the conductors, are kept in position and insulated from the outer tubes by means of porcelain spacing insulators. The length of feeder tube from the transmitter to each individual aerial wire in each aerial system is made to be exactly the same.

Thermionic tubes are used to generate the extra high-frequency oscillations and oil-cooled valves or tubes are employed on the main circuits in preference to water-cooled tubes because in short-wave work oil is easier to handle besides being itself an insulator.

The wavelengths in meters and frequencies in kilocycles used are:

For communicating with:

Canada	16.574	18,100
	32.397	9,260
Australia	25.906	11,580
South Africa	16.146	18,580
	34.013	8,820
India	16.216	18,500
	34.168	8,780

It is obvious that I am not even attempting to give anything like a complete description of these stations, nor have I said anything in regard to the buildings, power-plant, switchboards, rectifiers, controls, keying-systems, etc. To describe these important items and the receiving apparatus and stations together with an even brief account of the engineering and other difficulties encountered would require much more than one lengthy paper.

Other beam stations in England are nearing completion for the purpose of working high-speed services to corresponding stations erected by or in cooperation with the Radio Corporation of America in the United States and also with South America; and

similar stations have already been erected and are now carrying on a public service between Portugal and many of her distant colonies.

Mr. C. S. Franklin has to his credit the successful design and testing of most of the arrangements and devices employed, especially on the technical and scientific side; and Mr. R. N. Vyvyan for those on the engineering side. I hope they may soon be able

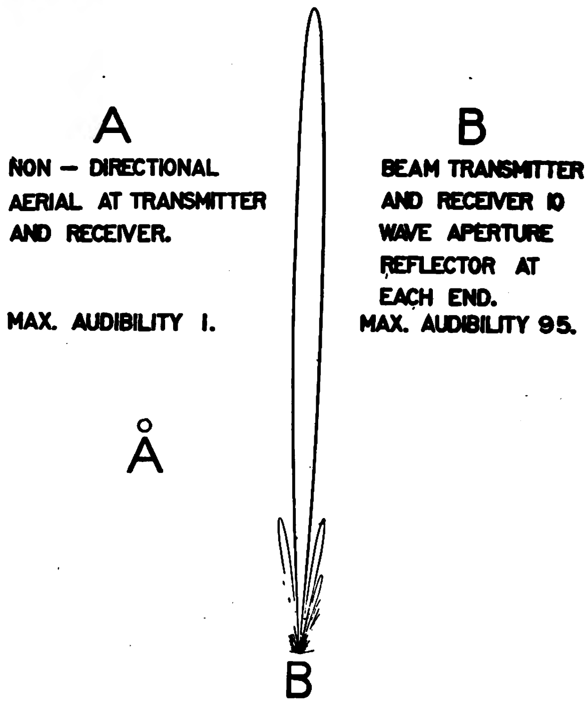

Fig. 7—Polar diagrams illustrating relative audibility at equal range and with equal transmitting power, assuming the wavelength and other conditions are the same.

to find time to prepare and publish a more complete description of these stations.

The commercial continuous working of these long-distance services over a period of many months has made possible the collection of observations of great scientific interest, and the carrying-out of further tests the results of which are, perhaps, quite novel in the history of long-distance radio.

One of them is in regard to electrical atmospheric disturbances generally termed static or "X's."

I think we all know that static has all along been the bugbear of radio, but one of the most salient facts that we have noticed in working long-distance services by means of short waves is that, particularly when receiving reflectors are employed, static has been generally conspicuous by its absence, and, when noticeable, the signal strength has mostly been well above the disturbance strength-level of static. Thunderstorms in the vicinity of the receiving stations occasionally interfere with working when they happen to be inside the angle of receptivity of the receiving reflectors, but not even then when they are at some considerable distance.

In working the high-speed receiving stations situated near Bombay throughout the whole of the East-Indian summer and during the monsoon period, interruption or interference with the signals received from England due to static has been of very rare occurrence.

I feel quite confident that our old enemy of static interference no longer exists as a serious hindrance to the working of high-speed radio as carried out by the beam system.

I fully realize that this is a bold statement to make, but I feel quite confident that I am right in making it.

The variations, or rather, the attenuation of signal intensity, now termed "fading" is the one, and I believe the only really serious difficulty with which we still have to contend.

Fading has been a marked feature of long-distance radio, especially when short waves are employed, and although in my experience fading appears to be worse on wavelengths between 200 and 1000 meters it has often proved to be serious on the very short waves now utilized by the beam system.

According to my experience, the use of reflectors has the advantage of diminishing the bad effects of fading. This is due, no doubt, to the very considerable increase of the average strength of signals obtained by the utilisation of the directional system which, thereby, increases the margin of readability of received signals enabling them to be still recorded or read through most of the fading periods.

I have been able to make some very interesting observations in regard to the phenomena of fading from the working of the short-wave beam system for world-wide communication of which England is now the center.

Fading has always been more frequent and more severe on

the England-Canada circuit than on any of the others. It may be noticed that our Canadian service is also our shortest distance service, that it is mostly across the sea and that the Canadian station is the one which happens to be nearest to the north magnetic pole.

Some interesting suggestions on the "Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism" have been set out in a paper by Mr. Greenleaf W. Pickard, published in the PROCEEDINGS of The Institute of Radio Engineers, Vol. 15, No. 2, of February of this year.

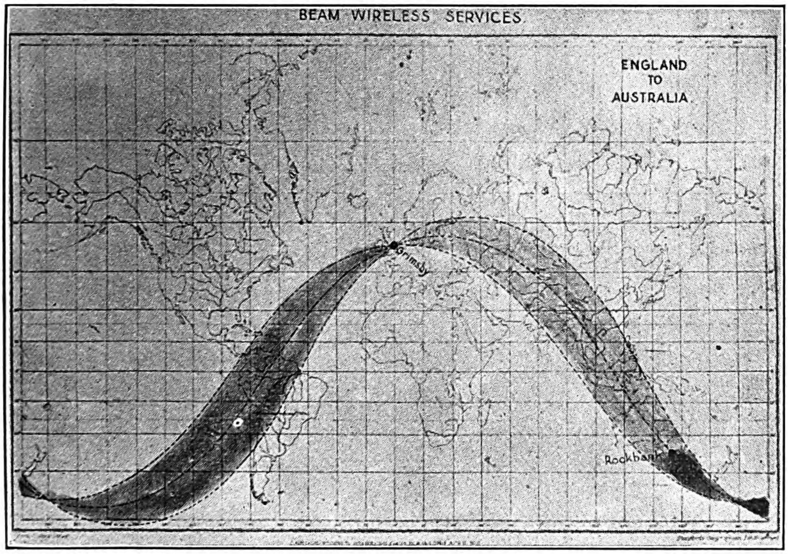

Fig. 8—Beam Services

It frequently occurs that when the Canadian communication fades out for some hours on end, the other services to Australia, India, and South Africa, which use similar wavelengths, continue working with undiminished efficiency. It has also been noticed that the times of bad fading practically always coincide with the appearance of large sun-spots and intense aurora-boreali usually accompanied by magnetic storms and at the same periods when cables and land lines experience difficulties or are thrown out of action.

We have also frequently noticed that during these periods signals could be received on a shorter wavelength than the one

usually employed, often on a 16-meter wave when a 26-meter wave would not come through.

As is now generally known, very short waves, of 16 meters and under, can be better received at long distances by daylight and in summer time than during winter or at night, and we also know that very long waves are not affected by daylight.

It may be that, on certain occasions during periods when sun-spots and auroras are prevalent, conditions due to the increased ionisation of the atmosphere at a certain height are prevalent,

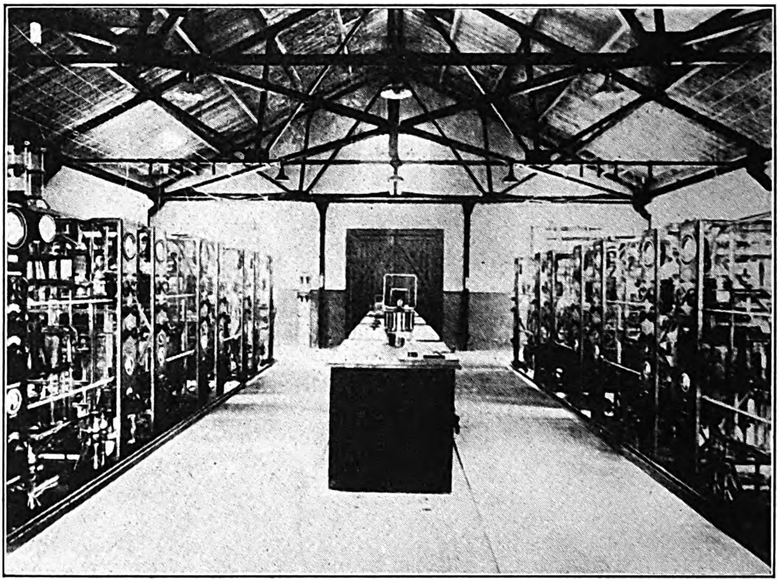

Fig. 9—Bodmin Transmitters. Left, Africa; right, Canada

resulting in the lowering of the ionised stratum which would produce an effect equivalent to what we might term "intensified daylight."

Professor Elihu Thomson in a lecture delivered in London during 1924[2] stated that it was his opinion that extensive auroral display was coincident with the existence of exceptional areas of disturbance in the sun, and he also referred to the probability under such circumstances of a decided elevation of charge, or potential, of the outer conducting layer 50 or 60 miles above the earth's surface.

[2] "James Forrest". Lecture delivered before The Institution of Civil Engineers in London on the 8th of July, 1924.

The phenomenon of fading is now being investigated by many workers, and progress is already being made in overcoming the difficulties which have been experienced in maintaining an absolutely continuous service between distant stations.

Time will not allow me to deal with this all-important subject which would necessarily require my referring to lengthy papers and reports and the publication of data which may still need experimental confirmation over protracted periods.

I shall, however, make a very brief reference to the subject of "skip distances."

This phenomenon, the study of which has been taken up since the advent of short waves, has been very carefully investigated

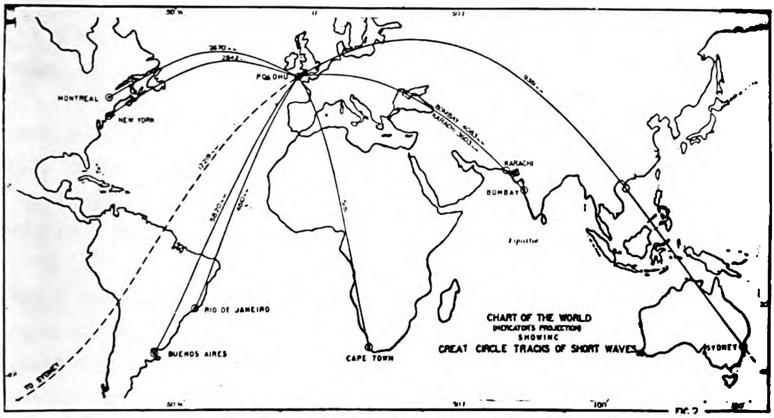

Fig. 10—Map of the World. Great Circle Tracks

by Mr. A. Hoyt Taylor and set forth and discussed by him in several admirable papers.

It is now, of course, well known that a wave of about 15 meters can be received with much greater strength and regularity during day time at distances about or over 5,000 miles than at distances of the order of a few hundred miles.

My experience, which I wish to record because it differs somewhat from the conclusions arrived at by Mr. Hoyt Taylor, is that when receiving on my yacht, even with land intervening between the two stations, there are no distances at which I have ever found zones of absolute non-reception, but that I have noticed zones where signals were weak and variable and where reception conditions closely resembled those prevalent at the normal ranges of the stations when fading conditions existed. In

these zones of weak reception, which more or less coincide with the so-called skip-distances, the received waves appear to be scattered in such a manner that direction finders fail to indicate any definite direction of origin or of propagation.

My observations were made on the S. Y. *Elettra* during a cruise which took place last August and September, when almost continuous observation was kept on the signals of eight beam stations situated at varying distances, both great and small, and working on wavelengths of approximately 16, 26, and 32 meters.

It may well be that the amount of energy radiated by these stations along the path of the beams was sufficient to give signals on our receivers even at distances over which reception could not otherwise have been detected.

Before concluding this address I would like to put before you a few considerations in regard to what I believe is the relative value of short waves versus long waves for long-distance radio.

We all know that space is becoming seriously congested over a very considerable range of wavelengths, and as we have only one medium of transmission for us all, it may be well to figure out roughly the probable number of possible wavelengths or channels which can be used without mutual interference.

If we assume that long waves may be classed between 5,000 and 30,000 meters, and short waves between 5 and 100 meters, then, by applying the basis of a rule proposed for the consideration of the International Radiotelegraph Conference at Washington, we find that 3,700 wave-bands or channels will be practicable and permissible for the short waves, but only 90 for the long waves.

This, of course, is rather a conservative estimate of the number of channels, but should narrower wave bands be adopted the proportionate permissible number of short wave-bands would bear the same proportion to the possible number of long waves.

But, in addition to this very great advantage for the short waves, they have a further one due to the possibility of restricting a large proportion of their power to within a narrow angle and also to the screening effect of the receiving reflectors which, by very greatly reducing the angle of receptivity and, thereby, minimizing interference, tends still further to increase the number of separate services which can be worked by means of these waves.

We should also not lose sight of the fact that very high speeds of working appear to be possible only if short waves are employed,

while, with the lower frequency of the long waves, speeds of the same order are quite unattainable.

I might, in other words, state that in regard to short waves there exists no theoretical reason why, with a frequency of 3,000,000, such as is the frequency of a 100-meter wave, the possible speed of working should not be 200 times greater than that attainable with a frequency of 15,000, which is that of the main transmitter of the high-power long-wave station at Rugby in England.

But long-wave stations, such as the one at Rugby, although

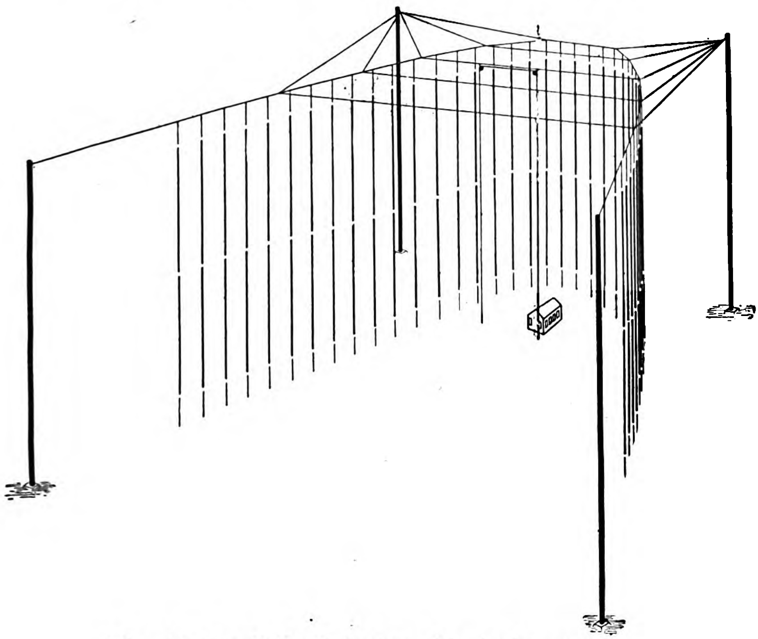

Fig. 11—Parabolic Reflector with Aerial at Focus

utilizing a power of over 500 kilowatts cannot even communicate at low speed with Australia for the same daily average number of hours as is possible by means of an efficient beam station, although the latter be using only 20 kilowatts of electrical energy.

There is also an interesting and economic feature in regard to long-distance services by means of short waves compared with cables which should not be overlooked.

With cables, the capital and maintenance costs of the cable itself increase in simple and direct proportion with its length,

but with short-wave radio it has been found that the capital cost of stations for communicating between England and Australia, over a distance of 10,000 miles, is materially less than the cost of stations for communicating with Canada over 2,500 miles: *i.e.*: over only a quarter of the distance, while, if anything, the service is better over the longer distance.

I have often been asked why it is that if these short waves are capable of covering the greatest distances without employing the beam system, I have always insisted on using it in all the stations my Company has established or is erecting for important long-distance services.

My reasons are that, as I have perhaps learned something in regard to the severe exigencies of present-day commercial telegraphy, I have realised that in consequence of fading and atmospheric interferences the signals obtained from such non-directive stations are rarely strong enough for operating the recording instruments necessary for high speed commercial services required between important far-distant countries.

Doubts have also often been expressed by some experts as to whether or not the reflectors and directive aerials used at the various beam stations were in fact fulfilling any useful object at all.

The tests over long distances have already shown that the use of beam aerials and reflectors at both ends results in a signal strength which, from careful measurements, Mr. Franklin estimates to be on the average about 100 times greater than that obtainable with non-directional transmitting and receiving aerials utilising the same power.

Now, since the increase of strength of the received signals rises in proportion with the square foot of the power of the transmitter, it is easy to estimate that in order to obtain signals 100 times the strength it would be necessary to use 10,000 times the energy and, hence, as the power supplied to the anodes of the tubes is 20 kilowatts it would be necessary to use the impossible and absurd power of 200,000 kilowatts with the ordinary all-round radiating and receiving stations to give the same average strength of signals at the receiving end, if suitable reflectors were not employed.

During my recent cruise on the S. Y. *Elettra* I had numerous opportunities of testing the strength of many of the beam stations over both short and long distances, and although on this occasion I did not carry suitable measuring instruments, I am quite sure

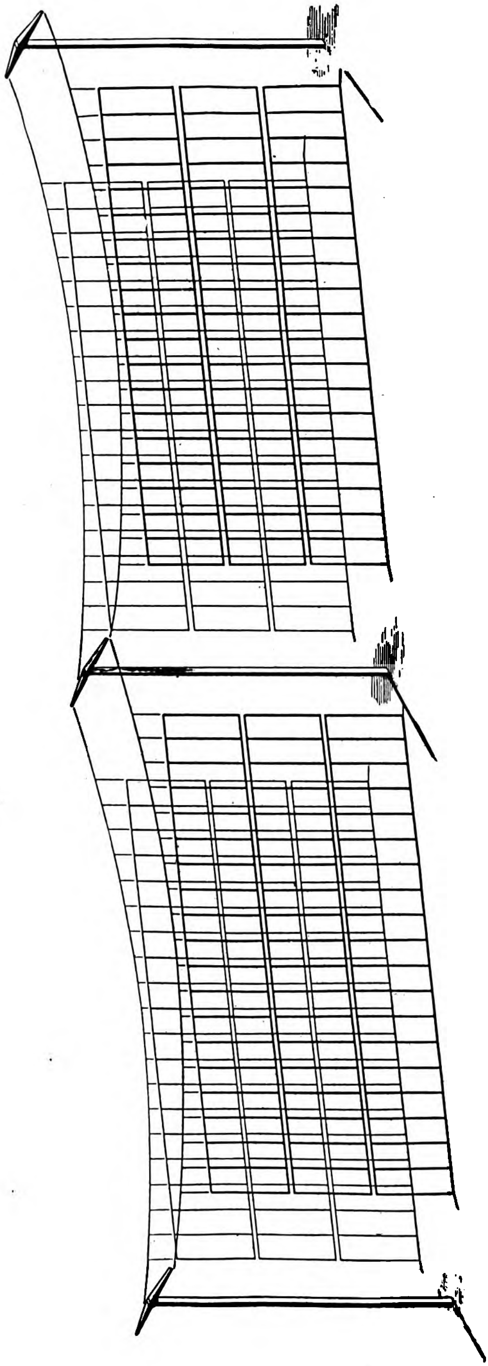

Fig. 12—Parallel Flat Aerial and Reflector

and satisfied that there is no possible doubt in regard to the very great increase in strength and reliability of the signals when these are received in the center or in the path of the beams as compared with those which can be received outside.

There exist, however, occasional periods, frequently coincident with fading conditions, when reflectors appear to be of no marked advantage. When such conditions prevail signals are seen not to arrive from any defined direction or angle but to be scattered to and from all directions, probably by the Kennelly-Heaviside layer, just as might occur from multiple reflection, or as in the case of diffused light.

This condition of scattering appears to prevail constantly at the so-called zones of skip-distances of each particular wavelength employed, and this fact may explain why certain observers in Germany and in America have found no trace of beam effect. If, as seems possible, their tests were made in the zone of skip-distances, the explanation appears clear.

These effects of the scattering of short electric waves have been carefully investigated by Mr. T. L. Eckersley of the Marconi Company who, I hope, will soon be able to publish a paper on the subject.

It is known that signals from the beam stations can be received at distant places far and wide and quite outside the angle of the beams, and it has also been suggested that this proves the inefficiency of the beam system.

It should, however, be remembered that a comparatively small amount of radiation escapes in all directions from the aerial and reflector system, and although this represents only a very small proportion or fraction of the energy contained in the beam, it is nevertheless capable of being detected, sometimes strongly, on sensitive receivers at very great distances in the same way as amateurs have shown us that the energy of a short-wave broadcasting transmitter, even when radiating only a few watts, can often be received and detected as far as the antipodes.

The same sort of effect occurs perhaps to a lesser degree even in the case of light when projected in a beam from a searchlight. Many of us may have noticed that it is usually quite easy to see plainly the beam of light projected by a searchlight and its glare even when the beam is not directed towards the observer.

The signals which can be received outside the beams are, however, rarely strong enough to work high-speed recording instru-

ments reliably, and as the messages which are sent in Morse code by the principal beam stations are normally transmitted at speeds approaching or exceeding one hundred words per minute, it is quite impossible to read them by ordinary aural or telephonic reception.

This affords a certain degree of secrecy not realisable by the older long-wave system.

Restriction of the angle of the beam and decrease of stray radiation outside of it appears, however, to be possible by increasing the dimensions of the reflectors in respect of the wavelength employed, and also by augmenting the number of wires in each grid. Further discoveries and the perfecting of design may also bring about this much-desired improvement.

Although the beam system is, in my opinion, still far from perfect, progress and improvement are continuous and the results already secured during many months of continuous working on a

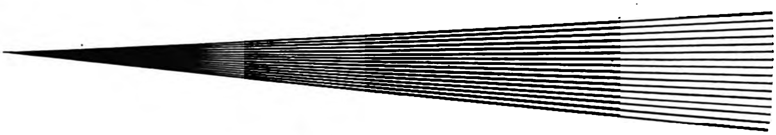

Fig. 13—The Beam

commercial basis across a variety of distances between so many different parts of the earth's surface have firmly convinced me that a good directional system is the system of the future for point-to-point radio communication over long distances throughout the world.

I have always felt that radio waves are far too valuable to be scattered and broadcast equally in all directions in a point-to-point service instead of being concentrated as much as possible on the station or group of stations with which it is desired to communicate.

The enormous increase in telegraphic speed is not the only advantage attached to the short wave-beam system: recent tests having fully demonstrated the adaptability of this system to radio telephony and also the ease with which it is possible to superimpose a commercial telephone channel upon high-speed telegraph services using the same system, as is now being done experimentally between Canada and England, thus obviating the cost of erecting separate stations for carrying on telephonic communications.

The commercial advantages of such an important development of the application of short waves are very great and the opening of the first multiplex telephone and telegraph service will certainly constitute an important new departure in the history of long-distance radio communication.

Much research work and the task of devising, designing, and testing the special arrangements and instruments used at the short-wave receiving stations have been successfully carried out by Mr. G. A. Mathieu of the Marconi Company.

Short waves are also beginning to show unhoped-for results in improving broadcasting and making it workable over great distances even during the hours of daylight. And directive methods, I feel confident, will soon be utilized for broadcasting by enabling programmes and speeches to be transmitted over large portions or sectors of America and to distant countries with much greater strength and freedom from interference than is possible with existing methods.

And, lastly, short waves cannot but enormously assist in rendering more practical the systems of picture and facsimile transmission including television, which are most likely to bring to an end the necessity for Morse code signal transmission on which is based telegraphy as we know it today.

In reviewing the progress recently made in the applications of short waves I may, perhaps, be forgiven if I say that it is with some considerable satisfaction and even pride that I am able to recall the fact, that when five years ago I last had the honor and pleasure of addressing you here, no practical use of short waves had yet been made and it was probably the first occasion on which the question of the urgent desirability of the study of short waves for practical radio purposes was publicly proposed and strongly recommended to the attention of experts.

I then stated that I considered we had perhaps got rather into a rut by confining all our researches and tests to long waves, because I felt that the study of short waves, although sadly neglected all through the history of radio, was still likely to develop in many unexpected directions and to open up new fields of profitable research.

The results obtained by amateurs by means of short waves do great credit to them, especially if we consider that most amateurs possess only limited facilities for experimental work. Their observations have frequently been of value in helping us to arrive at

a better understanding of the very complex phenomena involved, but care should be exercised in accepting their observations, especially when they concern what I might term negative results.

Some short time ago I read a statement by an eminent English authority that, according to amateur observations, the daylight range of a 100-meter wave did not exceed 200 miles, and for a wave of 50 meters, 100 miles.

These distances were very far short of those ascertained by myself and my assistants, but it may very well be that some of the observers did not possess efficient receivers or that the location of their stations was not favorable to reception, or that they lacked operating experience.

I have always found that for reliable comparative observations and deductions in regard to transmission over different and varying distances there is nothing so good as a receiving station installed on a suitable steamship.

On a ship or yacht, one has the advantage of using the identical aerial system, the same receiving apparatus, and the same observers throughout, and at all distances, and this is, I believe, a most reliable way of testing the behavior of these waves, especially at what I might call intermediate distances.

Looking back at our old difficulties, of only a few years ago, the ease and perfection recently achieved by radio, especially in regard to broadcasting, appears little short of miraculous. It shows us what can be done by the combination of a great number of workers all intent on securing improved results. And how many, who began as amateurs, have contributed in one form or another to this progress and to this success?

We are yet, however, in my opinion a very long way from being able to utilise electric waves to anything like their full extent, but we are learning gradually how to use electric waves and how to utilise space, and thereby humanity has attained a new force, a new weapon which knows no frontiers, a new method for which space is no obstacle, a force destined to promote peace by enabling us to better fulfil what has always been essentially a human need: that of communicating with one another.

A NEW METHOD FOR THE CALIBRATION OF AMMETERS AT RADIO FREQUENCIES

By

HERBERT C. HAZEL

(Radio Frequency Laboratory, Department of Physics, Indiana University)

Summary—*This paper describes the construction and operation of a thermionic vacuum tube which is designed for the measurement of radio-frequency currents. The input circuit consists of a filament whose cross section is small enough that the "skin effect" is negligible at the frequencies used. The filament is heated first by currents of known magnitude (at a low frequency) and again by radio-frequency currents to be measured. Electrons emitted by the heated filament are drawn to an anode sealed into the tube and a comparison of the resulting space currents indicates, if the necessary precautions are taken, the amount of current in the input circuit.*

DURING the last few years progress in methods of radio communication has given impetus to a study of the problems involved in the measurement of current and resistance at radio frequencies. While marked achievement has characterized the field of radio investigation in general, the method of calibrating ammeters for radio work has remained rather crude and inaccurate. Many problems arise in the design and construction of radio-frequency instruments due mainly to the fact that changes of current distribution with frequency may entirely change the effective resistance of a circuit. For measurements of extreme precision at very high frequencies a substitute type of measuring device, a thermionic ammeter, has been developed. This paper describes the new instrument with its auxiliary apparatus and associated circuits.

CONSTRUCTION AND OPERATION OF THE THERMIONIC AMMETER

If a radio-frequency current is sent through the filament of a vacuum tube, the temperature of the filament will be the same as that caused by an equivalent low-frequency or direct current, providing the filament is made of such small diameter that the "skin effect" is negligible. Now the rate of electron emission of a hot filament is a function of the temperature.[1] Consequently, if the potential gradient from the filament of the tube to a plate sealed into the tube is the same in the two cases, a radio-frequency current sent through the filament will produce the same space current in the plate circuit as would be produced by an equivalent low-frequency current through the filament. Obviously the

[1] Van der Bijl: "Thermionic Vacuum Tube."

current characteristic of the vacuum tube for low-frequency (say 60 cycle) filament current. This current may be easily measured very accurately by an instrument calibrated for use at the low frequency. By comparison of the resultant space currents as indicated by the meter E, in the plate circuit, the corresponding values of current in the input circuit can readily be determined.

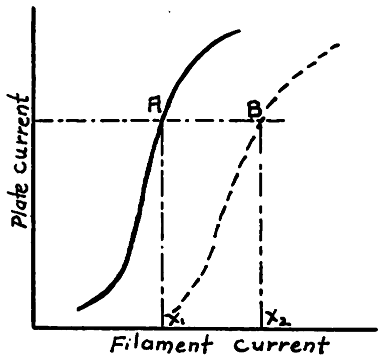

Fig. 2

If the radio-frequency ammeter used at A is in error, the characteristic curve resulting will not coincide with the curve obtained when accurately measured low-frequency current is used in the filament. For example, if the solid line of Fig. 2 be the curve for accurately measured 60-cycle current through the filament, and the dotted

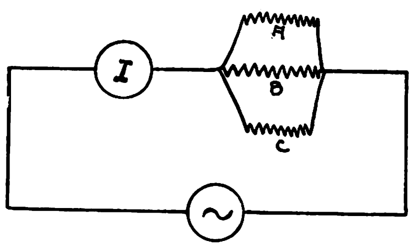

Fig. 3

line be the curve obtained with the radio-frequency current through the filament, then x_1 is the true value of the filament current which produced the plate current A. Then tracing across horizontally to the point B on the curve obtained with radio-frequency current at which the same plate deflection occurred, one sees that x_2 was the reading of the ammeter for the filament current which produced the plate current B. That is, x_2 is the

reading of the ammeter which corresponds to the true value of current x_1. The complete calibration curve may be easily obtained in this way. Of course one need not use any radio-frequency ammeter at all, but use the instrument just described to measure radio-frequency current directly.

The thermionic ammeter described above has several outstanding advantages. At extremely high frequencies the insertion

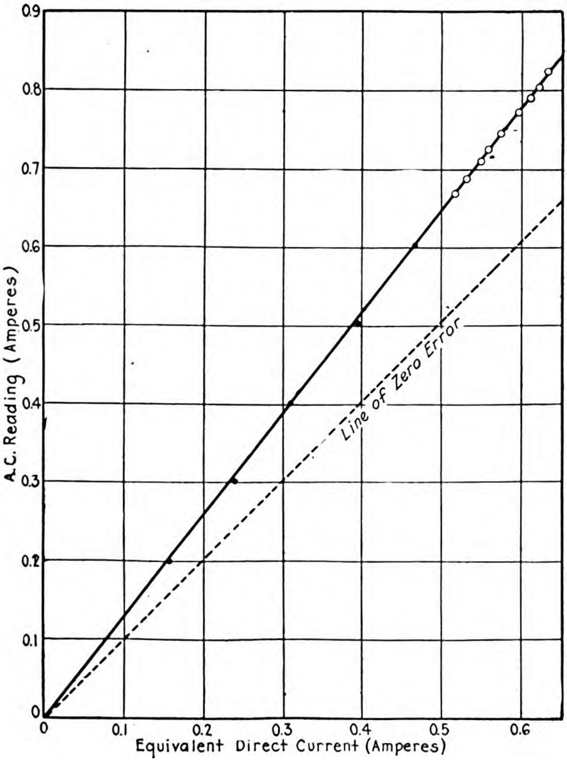

Fig. 4—Calibration Curve for a Commercial Ammeter (10,000,000 cycles).

of an ordinary ammeter in the circuit so profoundly changes the constants of the circuit that its use would be worthless. The specially constructed vacuum tube of the thermionic ammeter may be made of tiny dimensions with negligible inductance and capacity so that it may be used in circuits of extremely high frequency without appreciable error. The indications of the device, depending as they do on the fundamental property of electron emission of hot bodies, are practically instantaneous in action and are free from zero shifting. Since a given tube can, from the

nature of its functions, cover only an extremely limited range of current, and since the change of plate current with a very small increment of filament current can be made enormously large, the instrument is "microscopic" in its accuracy. Several tubes may be mounted on one base with a single milliammeter in a completely constructed ammeter offering opportunity for use over the desired range of magnitude, and yet retaining the extreme accuracy due to the fact that as incandescence approaches, the electron emission increases enormously. Several tubes may be used in parallel, the single milliammeter recording the sum of the plate currents of all the tubes. This would not be subject to the errors involved in putting a thermo-junction in contact with only one of a group of parallel wires neglecting the change in current distribution. That is, with reference to Fig. 3, I is not necessarily equal to three times the current through A (where A, B, and C are exactly alike) but is equal to the sum of currents through A, B, and C.

Calibration curves have been worked out by the new method for several Jewell, Weston, and Westinghouse radio-frequency ammeters at various frequencies, and have been checked by the calorimeter method. Fig. 4 shows one of these curves for a commercial instrument calibrated at a frequency of 10,000,000 cycles. The circles represent points on the curve obtained by the thermionic vacuum tube, and the solid line is the curve obtained by the calorimeter method. In this particular instrument the skin effect clearly predominates. As will be seen in the graph, the calibration curve has a greater slope than the line of zero error, and lies entirely above this line.

EXPERIMENTS AND OBSERVATIONS CONCERNING THE IONIZED REGIONS OF THE ATMOSPHERE*

By

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(Bell Telephone Laboratories, Inc., New York City)

Summary—*Experiments are described in which a virtual height of the reflecting ionized region was measured using time lag between impulses arriving over a direct, and the reflected path. The measurements were made on 57 and 111 meters. The height was ascertained only at night and the daylight hour before sunset. Movements of the reflecting region are plotted showing slow rises and rapid drops. The rising rate approximates 6 miles a minute, and the falling rate about 20 miles a minute. Multiple reflections were observed. Transmission measurement curves are given showing dependence of 16-7/8 meter signals on the night ionization, and the assistance that sunlight ionization can give. Experiments and curves are mentioned that show absorption to be one of the important factors causing poor daylight transmission in the wavelength region contiguous to 214 meters. It is pointed out that the absorbing region is below the refracting region and that the sky wave must make two passages through the absorbing region. A discussion is given to show that both electromagnetic waves from the sun, and β particles, must be assumed as producers of ionization to explain phenomena observed. By this theory the electromagnetic waves from the sun produce the ionization in the absorbing region, and part of the day ionization in the refracting region. This ionization is pictured as beginning at an altitude of about 16 miles and extending upward, and as experiencing diurnal and seasonal variations. The β particles produce, at an altitude higher than the absorbing region, further ionization which is the principal ionization at night due to the absence of the electromagnetic ionization. This ionization is pictured as occupying part of the same region as the electromagnetic ionization and being very irregular in intensity and position.*

THE great extent to which radio wave propagation is dependent upon atmospheric phenomena has been appreciated only during the last few years. That there was some relation has long been suspected. The first definite connection noticed was absorption, and it led engineers to look upon the atmosphere as a necessary evil. The phenomenon of transmission around the globe which could not be explained by any pure diffraction theory led to the first suggestions that the atmosphere might have something to do with the transmission as well as the absorption of waves. As investigations have progressed, it has become more and more evident

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* Presented at the meeting of the American Section of International Union of Scientific Radio Telegraphy, April 21, 1927.

that all radio transmission to appreciable distances is dependent upon atmospheric phenomena to a hitherto unsuspected degree.

With the knowledge that the atmosphere was of importance in the transmission of radio waves, the radio engineer naturally looked to the meteorologist for information in studying the subject. The time is rapidly approaching, however, when the reverse may be true—the meteorologist may be looking to the radio engineer for information in his studies. The additional tools and methods that the radio engineer provides may enable the meteorologist to obtain data concerning those parts of the atmosphere about which he has very little information at the present time.

In our short-wave experiments of the last few years several phenomena have been observed which it is felt have a direct bearing upon the subject of ionization in the upper atmosphere. Some of these things were observed in our transmission measurement work, a partial report of which has been published,[1] while others were observed in experiments interspersed with the transmission measurement work. Much of our work corroborates the experimental results of others and supports some of the hypotheses which have been advanced in explanation.

Reflected Waves

Among our special tests was the measurement by reflection of radio waves of the height of the reflecting ionized region. The work did not start with this object in view, but began as a study of quality distortion. That reduction of quality was caused by the variation of the carrier frequency had been shown by Bown, Martin, and Potter,[2] and it was desired to get some quantitative observations on the disturbing phenomena in the short-wave region.

The idea was conceived of deliberately wobbling a steady carrier and making an oscillographic record at a receiving station. The idea had scarcely taken form when it was realized that if the variation in received carrier amplitude is caused by interference between two beams that arrive over different paths a beat note would result whose pitch would depend upon the rate of frequency variation and the difference in path lengths. Our first tests were made in the summer of 1925. The results were sufficiently promising that it was decided to continue the tests with improved apparatus.

[1] Heising, Schelleng, & Southworth, I. R. E. Proc., Oct. 1926.
[2] Bown, Martin, & Potter, I. R. E. Proc., Feb. 1926.

The method used was as follows. Signals of the form of curve *a* in Fig. 1 were transmitted from our laboratory at Deal, New Jersey. The signals consisted of dashes of steady high-frequency power approximately 1/16 of a second duration with spaces of equal length between. The frequency of the antenna current was varied uniformly during the period that the current flowed as represented in curve *b* of Fig. 1. The frequency used was usually 2700 or 5260 (111 or 57 meters) kilocycles. The amount of variation Δf, during the period of a dash was between 800 and 4000 cycles. The signals were received at a field laboratory at Albertson

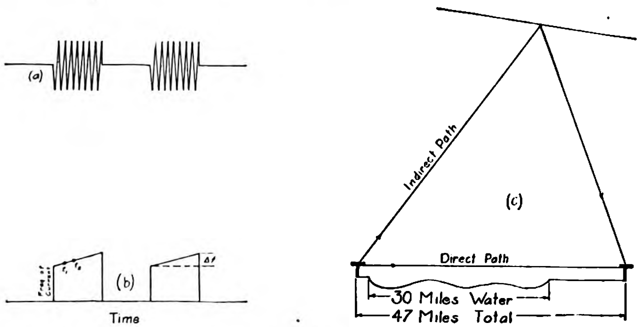

Fig. 1.—(a) Diagram showing type of antenna current pulses approximately 1/16 second long, and of varying frequency transmitted during "wobbler" tests. (b) Diagram showing how the frequency was varied at a uniform rate during the periods of transmission. Δf represents the frequency variation during a period. (c) Diagram showing probable paths with point reflection assumed for the indirect path.

near Mineola, L. I., a distance of 47 miles in a straight line as shown in Fig. 1, diagram *C*. The two paths expected for the two beams were the direct path, 47 miles, and the indirect path in which reflection or refraction occurred from the ionized region at some altitude. The wave radiated at time t_1 of curve *b* and traveling over an indirect path arrived at the receiving station at the same time that the wave radiated at time t_2 from the transmitting station arrived by the direct path. The two waves being different in frequency on account of their having left the transmitting antenna at different times produced a beat note. The frequency of the beat note was the difference of the frequency of the two waves, while timing, and measurements at the transmitter showed at what rate the frequency was varying during the period that the power was on. From these observations the difference between times t_1 and t_2 was ascertained, and this was the difference in the

times required for the waves to travel over the two respective paths. From this difference in time and the velocity of light, the apparent difference in path length and approximate distance of the reflecting layer was computed.

In Fig. 2, oscillograms VII, VIII, and IX show the transmitting antenna current. Except for a slight transient at the beginning of each dash the current was constant. To be sure that variations in current at the receiving station were not caused in the receiving equipment, oscillograms were taken at the receiving

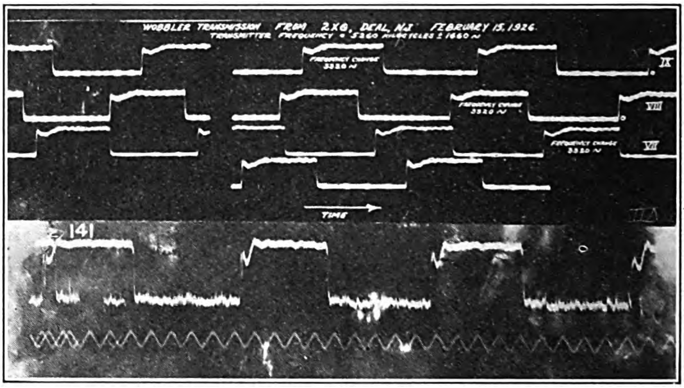

Fig. 2.—Oscillograms VII, VIII, and IX, showing variation in transmitting antenna current during "wobbler" experiments. Oscillogram 141, showing the received current at a time when no fading was occurring, Feb. 10, 1926, 3:10 P.M. The frequency was being varied over a 4000-cycle range during each period of this transmission. The rates of movement of the two films are not the same. The dip occurring at the start of each pulse was a transient in the transmitter (audio filter) and it is somewhat accentuated by the receiver square law detector.

station with no variation in frequency of the transmitting current. Oscillograms such as No. 141 of Fig. 2 were also taken at a time when fading was not present. This oscillogram shows no variation in amplitude beyond that produced in the transmitter, and by noise, even though the carrier frequency was varied over a 4000 cycle range during each dash.

The types of signals received at times when fading was present are shown in Figs. 3 and 4. The information concerning these figures is given in the captions and will not be repeated here. It may be remarked here that the heights given are virtual heights. The actual height is a matter of definition, since the waves penetrate a certain distance into the medium, and the actual velocity is not that of light while they traverse an ionized region.

The more carefully made records checked our earlier rougher records very well. We were not, however, entirely satisfied with the results because it was only occasionally that the beat frequency could be ascertained very accurately. Many oscillograms were secured showing beats too complex to allow of satisfactory analysis and it was thought some of them might have been caused by multiple reflections. It was therefore decided to adopt a different scheme so as to be able to measure not only the path difference but to ascertain how many paths there were.

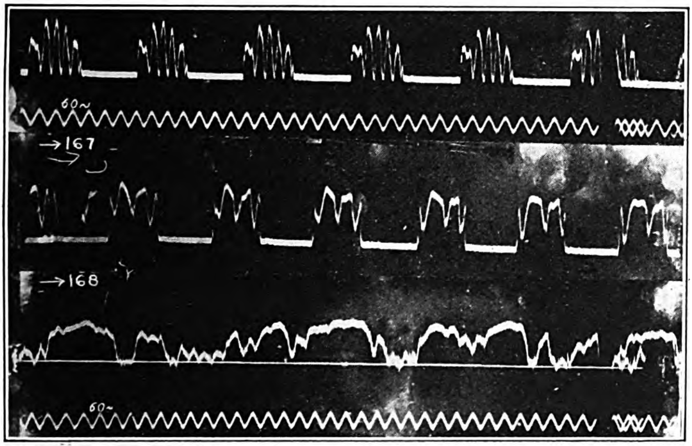

Fig. 3.—Wobbler signals received at Alberton, L.I., February 17, 1926. Average carrier frequency 5260 K.C. (57 meters).
No. 164 Δf=3320 cycles. Height of layer about 160 miles 4:08 P.M.
No. 167 Δf=1660 cycles. Height practically the same, 5:00 P.M.
No. 168 Rapid fading observed 5:30 P.M. while a constant carrier. was being transmitted.

The scheme adopted was somewhat similar to that used by Breit and Tuve.[3] Steady power of constant frequency was radiated in "jabs." These "jabs" of power were between 0.001 and 0.0016 second in duration, practically always the former. They were transmitted at the rate of about 60 jabs per second so as to provide between jabs a time interval that was long compared to the duration of the jab itself. When more than one path for the wave existed between the transmitter and the receiver, and the paths were different in length, a greater time was required for the wave to traverse the longer path than the shorter path. Under these conditions, for each jab transmitted, two or more jabs arrived at the receiving station at slightly different times. By measuring

[3] *Phys. Rev.*, Sept. 1926.

the difference between times of arrival of the jabs the difference in lengths of the paths was computed, and knowing the shortest path, or straight line distance, the apparent distance to the reflecting region was found.

Fig. 5 gives two oscillograms of actual transmitted jabs. Numerous oscillograms taken at the receiving station are given in Figs. 6, 7, 8, and 9. These are a representative lot among a large number that were taken. Fig. 6 is a group showing two paths quite clearly with the height of the layer of 221, 233, and 155 miles respectively, assuming the layer is nearly parallel with the earth's surface. The jabs arriving over the direct and indirect paths are

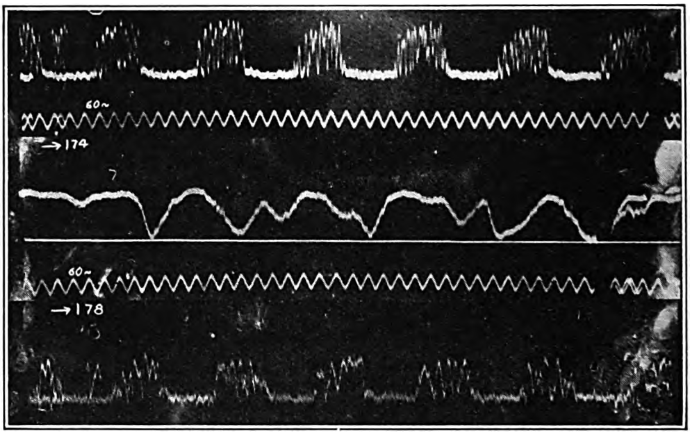

Fig. 4.—Wobbler signals received later the same day, February 17, 1926.
No. 171 $\Delta f = 3320$ cycles. Height, 250 to 300 miles; 7:48 P.M.
No. 174 Steady carrier alone, 8:06 P.M.
No. 178 $\Delta f = 1660$ cycles. Height, about 250 miles; 8:45 P.M.

quite distinct. They are distinct enough to be measured with considerable accuracy. The first film shows the signal arriving over the indirect path to be stronger than that coming over the direct path. The second oscillogram shows that conditions are reversed. The third one has the same characteristics as the second. It was taken with smaller amplification, and extraneous noises were much less. In some cases the oscillograms showed presence of noise about as great in magnitude as the received jabs but the regular re-occurrence of the received signals still enabled them to be picked out. In Fig. 7 is given a set of oscillograms showing more than one reflection from the ionized layer. One of them shows

three reflections while the others show two. The extra reflections show decreasing amplitude as would be expected.

Oscillograms were taken at times as shown in Fig. 8 to ascertain if there was a connection between fading and multiplicity of paths. All our records indicate that whenever fading occurred there was a multiplicity of paths. It was not, however, always possible to ascertain the number of paths or their difference in lengths because there were occasions when we did not get two or more distinct jabs but got a very broad one, as broad in time as two or three jabs following each other would be. This type of a received

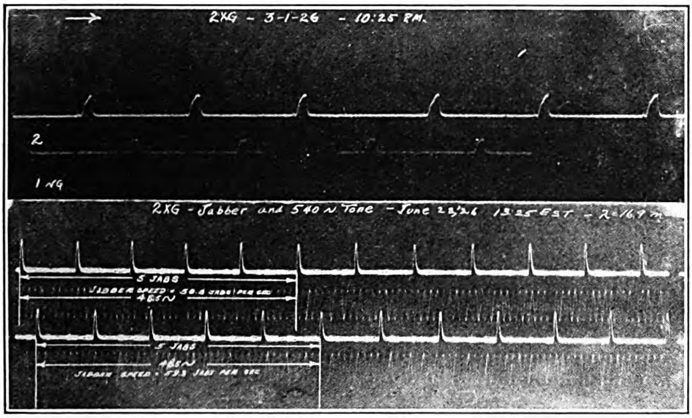

Fig. 5.—Oscillograms showing "jabs" of high frequency current in the antenna at the transmitting station.
March 1, 1926, 10:25 P.M., on 5260 kc. (57 meters).
June 23, 1926, 1:25 P.M., on 17800 kc. (16⅞ meters).

signal appears to be closely connected with the downward movement of the ionized region.

In the experiments reported by Breit he gives a probable height of the reflecting layer as under 141 miles for all observations. On only one occasion did our observations show as low a level as this. The heights which we found correspond to most of those secured by Taylor and Hulburt[4] by a totally different method. Most of our heights occur between 150 and 250 miles. We did find occasions when it was very much greater as indicated by oscillogram 267 in Fig. 9. In this case the height was of the order of 400 miles. The oscillogram given here happens to have been made on a different wavelength than the preceding ones, but that this was not the cause of the great difference in heights has been determined by other observations.

[4] *Phys. Rev.*, Feb. 1926.

Fig. 9, oscillogram 30 was taken at Dickinson, North Dakota at a still shorter wavelength than previous oscillograms, namely 16-7/8 meters, (17800 kc.) There are five paths apparent in this oscillogram with a sixth one developing. The straight line distance from Deal to Dickinson is 1510 miles. The signal represented by the fifth jab traveled a distance of 2000 miles greater than that traveled by the first jab, indicating that at the time of these experiments some of the energy traversed a path of at least 3500 miles.

As a rule reflection was observed during daylight only toward the end of the day. It is significant that that was also true of fading

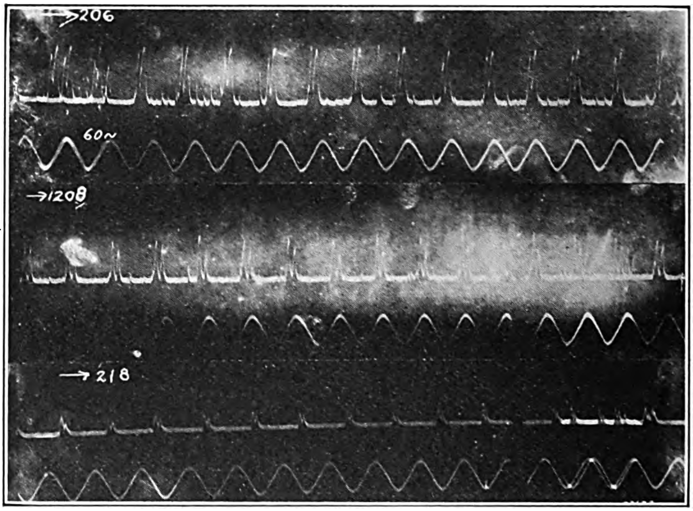

Fig. 6.—Oscillogram showing more than one path, taken at Albertson, L.I.
No. 206 Feb. 26, 1926, 9:19 P.M., 57 meters; height 221 miles.
No. 208 Feb. 26, 1926, 9:23 P.M., 57 meters; height, 232 miles.
No. 218 March 1, 1926, 4:41 P.M., 57 meters; height, 155 miles.

on wavelengths longer than 50 meters. On 67 meters fading would be first observed about an hour before sunset, and cease an hour after sunrise. For 111 meters the fading began very nearly at sunset and ceased as nearly after sunrise. If fading occurred during earlier afternoon hours during which observations were made it was too small to observe, and the same was true of reflections. This would be expected from our absorption observations and from signal strength measurements that are referred to later. It is estimated that in the middle of the day the reflected jab would be only a few percent of the magnitude of the direct jab, on 111 meters, and such a small signal would not show on the oscillograms.

Movements of the Reflecting or Refracting Ionized Region

In Fig. 10 are plotted curves from several series of observations showing the movement of the ionized layer. These observations were made at intervals of one or more minutes, usually depending upon the time required to change film holders and check up on oscillator adjustment.

The daylight or late afternoon observations secured indicate a level around 160 miles high. As evening passes, the general level rises. On March 1, 1926 the level rose from 160 miles about 4:41 P. M. to 200 miles about 8:00 P. M. and 220 miles about 10:15 P. M. On other occasions as may be observed from Fig. 10, it was higher. The highest level found was almost 400 miles at 8:34 P. M. March 5.

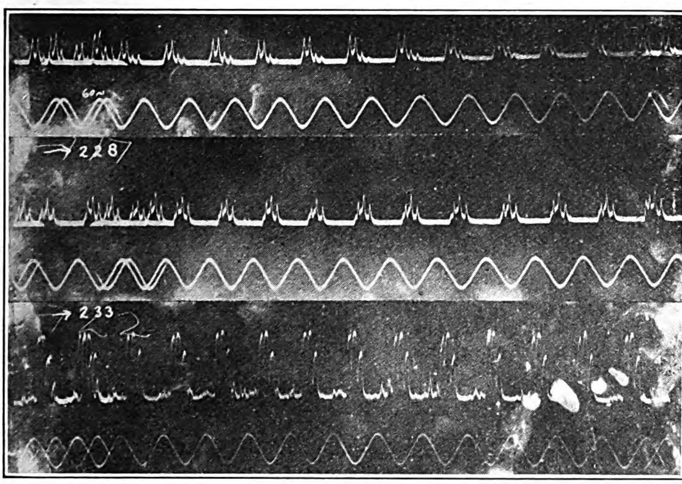

Fig. 7.—Oscillograms showing more than one reflection.
No. 226 March 1, 1926, 8:07 P.M., 57 meters; height, 200 miles. Path differences from direct path computed as 361, 762, and 1210 miles.
No. 228 March 1, 1926, 8:12 P.M., 57 meters; height, 200 miles. Path differences from direct path of 361 and 745 miles.
No. 233 March 1, 1926, 10:16 P.M., 57 meters, height 220 miles. Path differences from direct path of 394 and 835 miles.

The curves show an almost universal tendency to rise. The only exception is the one drawn between two readings taken at 10:58 and 11:00 P. M. March 3, where a drop is indicated of 45 miles in less than two minutes. Because the major part of the readings indicates a slow rise, and a few indicate a faster drop, it is thought that the level rises slowly and drops very rapidly, intermittently, with an average periodicity approximating once in about fifteen minutes for the occasions on which these observations were made.

The times at which multiple reflections occurred, indicating the layer was parallel with the earth's surface, were always times during which the level was rising and had been rising for some time. The multiple reflections were never observed just after a drop. This suggests a possible explanation. It would appear as though great masses of electrons are tossed into the atmosphere rather quickly and that as a result the level drops with accompanying turbulence and variation in density near the lower edge. Immediately thereafter, repulsion by the negative charge on the earth causes the entire mass to rise and the unevenness to disap-

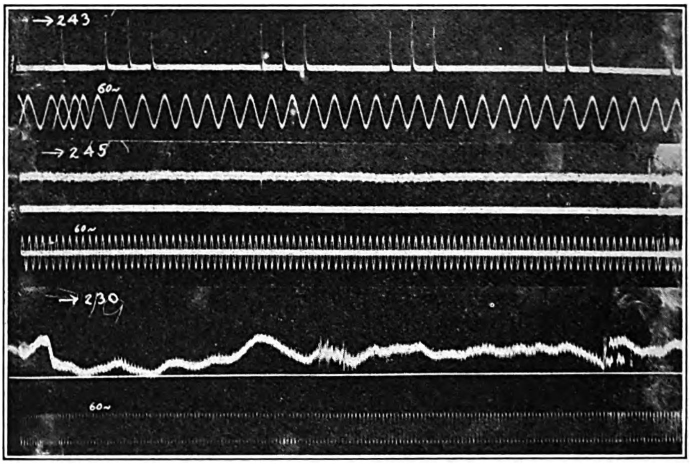

Fig. 8

No. 243 Type of signal received when no reflection occurred. The jabs were sent in groups on this occasion. March 3, 1926, 3:42 P.M., 57 meters.

No. 245 Showing absence of fading on the same afternoon as No. 243; 3:49 P.M. when a steady carrier was transmitted.

No. 230 Fading occurring when steady carrier was sent on the occasion of oscillograms of Fig. 7. March 1, 1926, 10:09 P.M., 57 meters. The 60-cycle hum in the current should not be confused with the fading.

pear. After the mass of electrons has moved upward for a few minutes, the disposition approaches uniformity in a horizontal direction and we get the phenomenon of multiple reflection. Immediately after the arrival of another mass of electrons, the layer drops with a turbulent movement again and the multiple reflection disappears. There are times as stated previously when no height can be computed due to the absence of two distinct jabs on the oscillogram, but where a broadening of the single jab occurs to such an extent that there is clearly a time lag involved. It is

thought that these observations occurred while the layer was descending. At 8:11 P. M. March 1 between two multiple path observations a two-path observation was secured in which the second path jab was a broad large jab indicating a turbulent condition and leading us to believe that a mass of electrons had arrived in the few minutes previous and lowered the layer level. The lower level measured at this time as compared to the previous and subsequent times strengthens this belief.

The heights as given in this paper are virtual heights. They are calculated on the assumption that ordinary reflection takes place and that the layer is parallel to the earth's surface. This latter is not necessarily true as the layer may be at some angle as represented in Fig. 1 in which case "distance to the layer" would be the more proper phrase. The distance between the transmitter and receiver was small in comparison to the distance to the layer, so that any error that would occur in computation due to varying direction is very small. As regards the questions involved in assuming regular reflection at the ionized layer, reference is made to the paper by Breit and Tuve where the matter is discussed in some detail.

Long Distance Transmission Observations

The experiments just described were all made with wavelengths of 57 or 111 meters. Further light on the phenomena connected with the ionized regions is obtained from field strength measurements on long distance transmission.

In Fig. 11 are given diurnal curves of average field strength on 16-7/8 meters as measured simultaneously at three points during the first two weeks of June 1926. The measurements were made at Dickinson, N. D. (1510 miles), Seattle, Wash. (2420 miles), and New Southgate, Eng. (3560 miles). The transmissions were made from Deal, N. J. It is a significant fact that though the two most widely separated stations are over 122 degrees different in longitude and on opposite sides of the transmitting station, the diurnal curves are very similar. In the paper previously referred to[1] there was mentioned a depression in the transmission curves and surfaces which appeared around midnight on wavelengths of 45 meters and under (page 627). This depression also appears in data of other observers. It is very pronounced in these three diurnal curves. Since the time to which the curves are plotted is Deal time, it is apparent that the night time conditions near the transmitting station predominate in producing this pronounced dip in the

curves. Mr. Schelleng has suggested a possible cause for this phenomenon which will be more easily understood by reference to the following table:

Table showing the distances from various places to that region in the sky to the north at midnight (at each place, June 21) where sunlight may be found, assuming no refraction by the atmosphere. Also height of that region above the earth.

	Distance	Vert. Height
Deal	403 miles	363 miles
Dickinson	232 "	218 "
Seattle	214 "	202 "
New Southgate	137 "	132 "
Most northerly point on great circle path between Deal and New Southgate	105 "	102 "

The table shows that at midnight in June at New Southgate, Dickinson, and Seattle, the distances to the region of perpetual sunlight are not very great. They are of the order of magnitude or less than most of the heights shown in Fig. 10. Since a short wave such as was used in these tests penetrates further into the ionized region than the waves used in the tests of Fig. 10, it is reasonable to say that the distances to the region of perpetual sunlight at these locations is less than one would expect to find for the effective height of the ionized region at night on this wavelength.

At Deal the distance to the region of perpetual sunlight at midnight in June is much greater than the distances at the receiving stations and greater than the heights shown in Fig. 10. The night at Deal produces a pronounced effect on the transmission in both directions. It is therefore apparent that *sunlight produces some of the ionization in the refracting region.*

As corroborative evidence of the above, it may be stated that similar curves taken at New Southgate, England, at other times of the year when the distance to the region of perpetual sunlight is large, show as pronounced an effect of the night at the receiving station as is produced by the night at the transmitting station. Signals are received practically only during mutual daylight hours. It appears as though *sufficient ionization for satisfactorily refracting 16-7/8 meters is produced only by sunlight.*

Another conclusion to be drawn from these curves and data is that a wave such as 16-7/8 meters travels to an altitude of between 200 and 400 miles on its passage from Deal to the receiving locations mentioned. Curves for 22 1/2 and 33 3/4 meters corresponding to those of Fig. 11 show that the longer waves travel at progressively lower altitudes since night effects of the receiving locations become progressively more noticeable. This agrees with the refraction theory as well as with the observations of Taylor.

The significant deductions from these experiments can also be

stated as follows: (1) Sunlight, probably the ultra-violet and gamma rays, produce ionization in the atmosphere which assists in the refraction of radio waves. (2) The refraction at night on 16 to 25 meters occurs well over 100 miles up, probably over 200 miles up, but probably not over 400 miles up. (3) After midnight in the absence of sunlight there may be insufficient ions to refract satisfactorily the shorter waves though other data show that there are plenty to refract longer waves (67 or 111 meters.)

Absorption Observations

Some further observations that may throw light on the factors affecting transmission are contained in the I. R. E. paper previously referred to.[1] Fig. 12, which is taken from that paper, shows ratios between average night and average day signals as a function of frequency. It is to be observed that the ratio of night to day signal strengths decreases as the frequency is increased, and that finally the day signal will be the stronger, depending upon the distance and frequency. On account of the shape and positions of the curves, it appears necessary to assume that absorption plays a large part in the phenomenon, and the observations admit of a simple explanation on this basis.

We may take, for instance, the case of transmission from Deal to Fairfax, Va. or Emporium, Pa. The distances are 215 and 243 miles, respectively. The location of either of these stations is such that the ground wave reaching it was reduced to much less than five per cent of the night signal strength no any wavelength measured. The received wave was therefore practically entirely an overhead wave. During the day the strength of signal received on such a wave as 67 meters will depend, among other things, upon the height of the reflecting region, as it affects both the angle at which the radiation must leave the vertical transmitting antenna, and the actual length of path traversed. At night the reflecting region is higher. The increased height will lengthen the path and increase the angle at which the signal leaves the antenna. Both effects tend to reduce the strength of the received signal. However the signal actually received on 67 meters at night is the stronger. This discrepancy is accounted for by absorption. However, there are wavelengths, such as 45 meters, for which the day signal is actually stronger. Hence forty-five meters is a wavelength for which at these distances the absorption by day is less than the reduction by night due to increased distance and greater angle of the wave leaving the antenna. For greater distances than these

two places mentioned, the same rise in height of the refracting region would produce a smaller change in path length and a smaller change in angle of leaving the transmitter so that a smaller reduction in night over day signal would tend to occur and less absorption would be sufficient to neutralize it. The absorption decreases with wavelength in the region under consideration and hence it would be expected, as is found, that for greater and greater distances, the frequency must be increased more and more to secure better day than night transmission. There is, of course, a limit in this direction due to variation in refraction with frequency, to

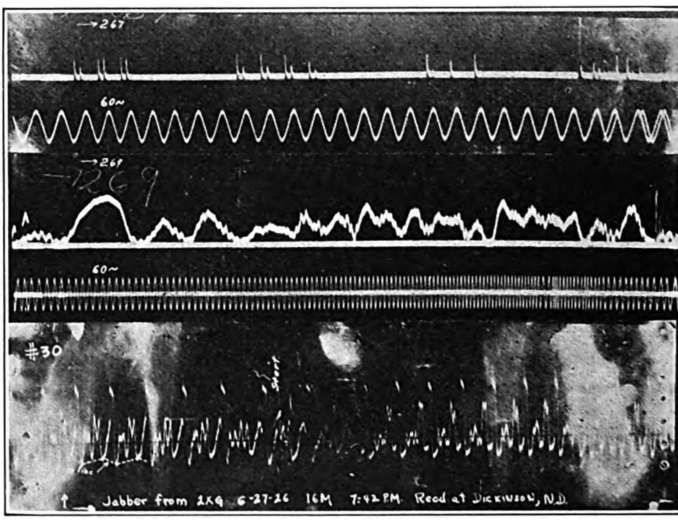

Fig. 9

No. 267 Two paths with the ionized layer very high. March 5, 1926, 8:39:30 P.M. 111 meters; height, 397 miles.

No. 269 Fading on the same evening, steady carrier radiated. March 5, 1926, 8:42:30 P.M. 111 meters.

No. 30 Multiplicity of paths observed at Dickinson, North Dakota, on June 27, 1926, 7:42 P.M. E.S.T. on 16⅞ meters. Paths are about 800, 1300, 1600, and 2000 miles longer than the most direct path, with a path 3000 miles longer developing, if jabs between the large jabs are delayed from the large jab just preceding. In this oscillogram the "zero line" is variable as a transformer was used.

the height of the refracting region, and to the curvature of the earth.

Discussion

That sunlight affects the transmission of radio waves has long been known. For the longer of what are called short waves, such as 67 meters and over, the signals are strong during the night

hours, the signals disappear very quickly when sunrise occurs at either end of the path and they appear just as quickly after sunset has passed the last of the communicating stations. The sunlight appears to hinder their travel, though it aids the shorter waves.

The short wavelengths longer than 50 meters usually have fairly uniform transmission during dark hours. They do not experience the falling off after midnight that is found in the case of the still shorter waves. The 16- to 50-meter waves require a greater

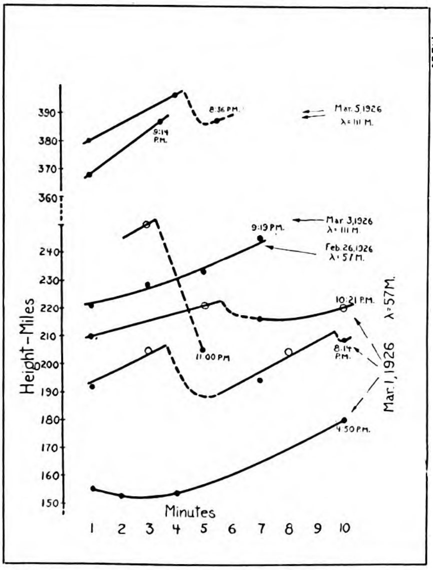

Fig. 10.—Variation of Height of the Region with Time. Circles and dots are observations. Circles are observations in which multiple reflection occurred.

electron density or density gradient for refraction than the 50–150-meter waves and the falling off in signal strength after midnight is probably due to a reduction in the number of electrons to the point where the shorter waves only are seriously affected.

Fading and reflection, which have been shown to be coexistent, were not observed at Albertson on 57, 67, and 111 during most of the day. Fading appeared on 57 or 67 meters before sunset, often as much as an hour and a half, but would not appear on 111 meters until practically sunset. Reflection was observed before sunset on 57 meters on several occasions, while on 111 meters it could not be found. Since a smaller electron density or density gradient is needed to refract 111 meters than 57 meters, the failure to get the

reflection on 111 meters when it was found for 57 meters could not be ascribed to failure in reflection, but to absorption which is closely connected with sunlight. It is known from other experiments that both fading and reflection would be observed in these experiments if the receiving station were at such a distance that more amplification was necessary for the ground wave, or if amplification had been "cut in" at the instant the reflected jab was expected.

Diurnal transmission curves of field strength indicate a daylight influence. On the longer of short waves where one can be sure the electron density or density gradient seldom falls to too low a value to refract sufficiently, the night average intensity of signals is surprisingly constant while the day intensity has a minimum around noon. The 111 meters suffers more attenuation than 67, and 67 more than 45 meters. At night the signal strengths on the three may be closely the same, while during the day they are widely different. Some of these characteristics appear in the curves given in the paper previously referred to by Schelleng, Southworth, and the writer. The shapes of the curves strongly suggest that these phenomena are due to absorption.

Since the reflection of 57 meters before sunset indicated the height of the layer was 150 miles or more at times when no reflection on 111 could be observed, the question arises as to whether the absorbing of 111 and refracting of 57 meters are accomplished by the same ionized region.

The effect of free electrons in the atmosphere upon radio waves has been discussed by Eccles,[5] Larmor,[6] and Nichols and Schelleng.[7] The correlation of this type of phenomenon with other radio transmission and cosmic data would appear to require the postulation of two distinct regions in which refraction and absorption occur. The free electrons may be produced by an ionizing agent operating on the gases of the atmosphere liberating free electrons, or they may come from an outside source thereby producing an excess of electrons over what are needed to render the gases neutral. It is desirable at this time to emphasize a point not discussed in these papers, but yet obviously present and underlying the discussions, which is of fundamental importance in interpreting these experiments if we hold to the idea that free electrons are responsible for the phenomena. The point concerns the possible locations of the regions in which absorption or refraction

[5] *Proc.* Roy. Soc., June 1912.
[6] *Phil. Mag.*, Dec. 1924.
[7] *Bell Tech. Journal*, April 1925.

can occur. The assumption necessary to any theory of refraction by free electrons is that the mean free path be so long that the electrons can absorb energy from the passing wave and then reradiate it all in a slightly different phase. If the mean free path is small, the electron may strike a gas molecule or ion before it has reradiated all the energy absorbed, and the remaining energy will be lost to the advancing wave. There will be a much greater chance of this loss in energy where there are greater numbers of gas molecules present. As pointed out by Nichols and Schelleng, the greatest loss of energy per electron will occur at some certain

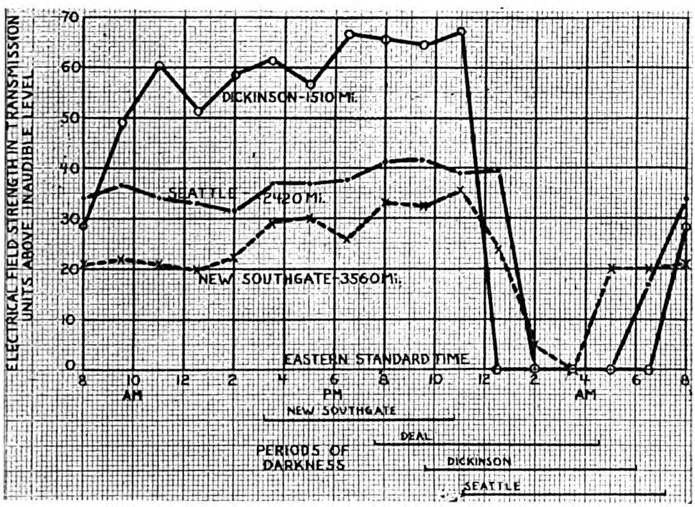

Fig. 11.—Diurnal Curves Showing the Average Field Strength of Signals on 16⅞ meters as Measured Simultaneously at London, Seattle, and Dickinson—June, 1926.

density of the atmosphere, it being expressed as the place where, (for no magnetic field, or for the electric vector parallel to the magnetic field) the electron collision frequency equals 2π times the radio frequency. For other wave components in a magnetic field it is slightly different from this. For altitudes below that at which this required condition obtains, the absorption quickly becomes small or negligible and the same occurs above this altitude for most components. Below this maximum absorption altitude which in this paper is termed the "absorption region," the electrons do not have a chance to move under the influence of a wave long enough to absorb and reradiate much energy before they strike gas molecules, and therefore there can be no refraction, while sufficiently above this region they may seldom strike molecules

and there will be little absorption. In the desirable refracting region the collisions will be so small in number as to entail little or no absorption. From these considerations it may be deduced, therefore, that *the refracting region is always above the absorbing region and the "overhead" or "sky" wave will always pass through the absorbing region twice in its path from one point on the earth's surface to another.*

With this picture of the absorbing and refracting regions, it becomes increasingly clear why the very long and very short waves can be used for long distance communication while the region in between is useless during daylight hours. The unsatisfactory results attending efforts to ascribe transmission around the curved surface of the earth to diffraction, and the effect of a conducting earth, have left the alternative of ascribing such transmission to refraction above the earth's surface. All long distance communication must therefore be dependent upon refraction, and although all waves from 15 to 10,000 meters are satisfactorily refracted, only those waves within this region are satisfactory for long distance communication if they can make their two or more passages through the absorption region without undue attenuation. Nichols and Schelleng show definitely in their paper that one effect of the earth's magnetic field is to reduce enormously the absorption that occurs on the longer waves—such as 2000 to 10,000 meters. The effect of the magnetic field on the absorption of waves under 100 meters is small, but the mass of the electron begins to play the same part here as does the magnetic field for long waves. As a result, it will be found that the absorption constant is small for the shortest usable waves, that it is increasing rapidly as the wavelength is increased, and that it passes a maximum slightly above 214 meters and decreases again in the longer wave region for some components of the wave. The failure of the middle region to be of any use in long distance communication during daylight hours is thus easily explained purely upon the basis of electronic absorption.

With this picture as a basis, it also becomes possible to explain many phenomena including those described in this paper, and to show that they point to the existence of two ionizing agencies instead of one, both agencies having been suggested many times but not indicated previously as being equally necessary.

The locations of the regions of the atmosphere where absorption and refraction occur can be deduced reasonably well from our knowledge of the structure of the atmosphere. Using the table

given by Chapman and Milne[8] for the pressures and mean free paths of gases in the atmosphere, the altitude at which maximum absorption occurs for 50 meters is computed as about 40 miles. For longer waves it is higher, being around 15 or 20 miles higher for 3000 to 10,000 meters. At 50 meters the depth of the absorbing region is fairly narrow, being about 15 miles deep so that the refracting regions for 50 meters could be considered as beginning

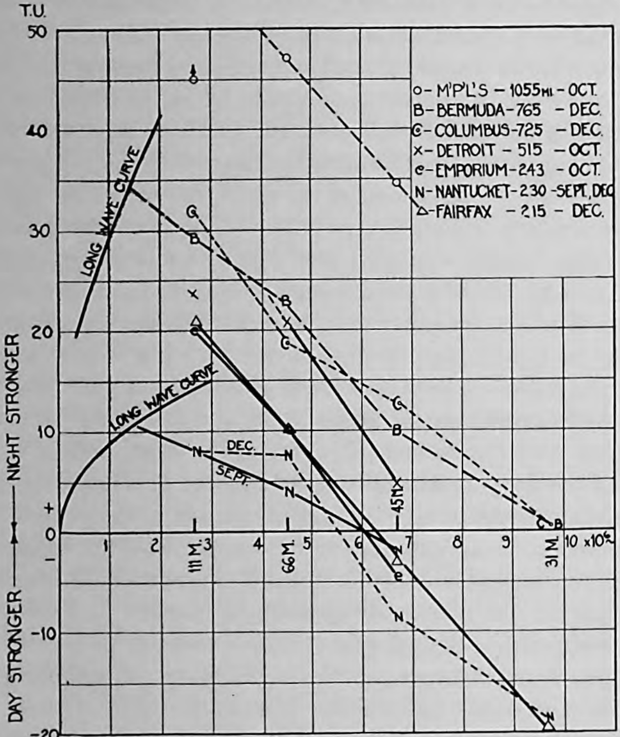

Fig. 12.—Ration of Average Night to Average Day Signals as a Function of Frequency. (Tests made in 1925).

around 50 miles up and continuing on up indefinitely. For shorter waves than 50 meters, the absorbing and refracting regions are lowered a few miles while for longer waves they are slightly higher.

An item of importance is the electron distribution and its cause. The close connection between poor long distance transmission on the wavelengths contiguous to 214 meters and daylight indicates that electro-magnetic rays from the sun cause the ionization in this region. It appears very improbable that the β particles from the sun cause the ionization. Chapman and Milne have estimated

[8] *Quarterly Journal* of the Royal Meteorological Soc. V. 46, 1920.

that the height of the region of maximum ionization by high velocity electrons from the sun is about 35 miles. While this is below the computed altitude of the absorbing region, it would not appear to be the cause of poor daylight transmission since the electrons cannot approach the earth in straight lines and produce ionization uniformly over the sunlit side, but must come along magnetic lines of force. They can arrive only near the polar regions except under most unusual conditions. The phenomenon of absorption should be confined to that part of the earth's surface at which the gaseous ionization occurs unless forces are present which cause the free electrons to migrate to all latitudes *before* rising under the influence of the negative charge on the earth, to an altitude above the absorbing region. Such migration from the polar regions would be expected on the dark side of the earth as well as on the sunlit side and produce absorption at night. The number of electrons reaching the tropics would be less than the number reaching the temperate zone thereby producing a variation in absorption with latitude. The absorption would be expected to vary widely with variation in the number of β particles arriving. Such expected phenomena have not been observed. Besides, it is not evident how such a hypothesis can be made to fit the following facts. The absorption on 67 meters or 111 meters is maximum about the middle of the day. It falls off toward sunset. It is small enough an hour before sunset due to obliquity of the sun's rays so that fading is observed on 67 meters and on 57 meters, and the existence of multiple paths can be determined for 57 meters. After sunset the absorption on 111 meters disappears so quickly and completely that this wave may be used for transmission up to several thousand miles. Long distance signals on 111 meters disappear quickly at sunrise, and on 67 a short time afterward. Only a cause which disappears at sunset and reappears at sunrise, without a hold-over effect or time lag period of any magnitude can be satisfactorily assumed. The ionization of the atmosphere by ultra-violet or gamma rays appears at present to be the most probable cause of free electrons in the absorbing region.

Computations made to determine the probable ionization by waves from the sun and based upon known penetration constants for gamma rays in various substances, and upon Chapman and Milne's structure of the atmosphere, indicate that the maximum number of electrons would be liberated in a given time at an altitude of 16 miles. The number produced would decrease very rapidly below that point, and decrease much more slowly above it.

The ionization produced by this means thus extends to a sufficiently low altitude to include the absorbing region.

In the refracting region conditions are more complex than in the absorbing region. It is more difficult to fit all experimental facts to a given hypothesis. In general, it has been assumed that the refracting region was "the ionized layer" where the free electrons were produced either by electromagnetic radiation or β ray impact, or by the β ray electrons themselves. The experiments described in this paper appear to support the hypotheses that refraction is assisted by, and at times is entirely due, to electrons entering the atmosphere.

A few of the points to be kept in mind may be listed as follows:

1. In the refracting region the electron collision frequency must be small, that is, the atmosphere must be very rarefied.

2. The amount and direction of refraction depends upon the electron density, the electron density gradient with height, and the magnetic field gradient with height. An electron density gradient appears to be the most necessary condition though an enormous change in refraction may occur in certain wave components by a change in electron density only.

3. The ionized condition lasts throughout the night sufficiently well to refract downward wavelengths of 67 meters and longer.

4. The conditions of ionization may change sufficiently during the night to affect seriously the refraction on wavelengths shorter than about 50 meters.

5. The refraction is noticeably increased on the shortest waves by sunlight. If the sunlight can operate on the refracting region at a time when refraction is not good, the refraction may be changed from an unusable to a usable value.

6. An electron density of 10^5 appears necessary for refraction back to earth of 50 meters and $10^{5.5}$ for 16 meters (From Nichols and Schelleng's approximate formula).

7. The ionizing conditions after sunset are not constant nor do they change at a uniform rate, but actually vary enormously. This is very noticeable as regards the height of the region.

8. A wide variation in the apparent height is observed from night to night, the values obtained being anywhere from 100 to 400 miles.

If it is attempted to explain the above phenomena on a hypothesis involving electromagnetic radiation only, two physical facts stand out as difficult to explain. The first is the persistence of the ionization after sunset. The second is the variation in height which

occurs after sunset, both from minute to minute and from day to day. Take first the matter of the persistence of ionization after sunset. It is usually suggested that the time of recombination would be so long that the ionization, though decreasing, could last throughout the night. Computation of the falling off of electron density after sunset on the basis of recombination is very difficult on account of the lack of sufficient information. However, an approximate computation can be made. Since the reflection of waves of 60 meters and longer is always observed hours after sunset, the electrons must still be present in the refracting region at this time to the amount of 10^5 per cc. The electron collision frequency and number of gas molecules per cc. are respectively 2500 per second, and 5×10^{11} (100 miles high) and 100 per second and 2×10^{10} (200 miles high). A calculation based on the assumptions that all collisions of electrons with ionized gas molecules result in combination, that the attraction of the positive ions and electrons and the effect of neutral molecules is ignored indicates that a period of less than 30 minutes will suffice for a reduction of electrons from 10^7 to 10^5 per cc. at these altitudes. The time would be longer if the temperature is lower than that assumed by Chapman and Milne, -54 deg. centigrade, and shorter if the temperature is higher.

The other phenomenon which it is difficult to account for is the rapid variation in height of the region from minute to minute, and to a lesser extent from night to night. The layer was observed to rise at a rate of about 6 miles per minute, and fall at a rate of about 20 miles per minute with a period approximating fifteen minutes. Such movements in the ionized region after sunset if the electrons are produced by electromagnetic radiation would require that rays of varying strength approach the earth on the dark side, or that great electric fields cause the uncombined electrons to migrate up and down in the air.

The difficulties attending the fitting of facts to the ultra-violet light hypothesis make it desirable to give some attention to a possible explanation by the β particle hypothesis. Many of the experimental facts are easier to explain on this basis. To begin with, one must differentiate between the ionization caused by collision as the arriving β particles are brought to rest, and the effective ionization caused by the presence of the arrived electrons alone after the original gaseous ionization has recombined. The β particles cannot approach the earth in straight lines over the sunlit side, but are constrained to follow down along the magnetic field into the polar regions. It has been computed that they produce the

greatest ionization at around a 35-mile altitude. The ionization by collision therefore can occur only in the polar regions and not uniformly over the projected surface of the globe, so that the presence of free electrons in the refracting region over other than the polar regions must be accounted for by movement from these latter regions. The gaseous ionization at the ionizing level will quickly disappear, but the excess of electrons occurring due to the arriving β particles will cause a space charge to occur which can cause a migration of electrons in all directions. The effect of the atmosphere, the earth's magnetic field, the negative charge on the earth, and the rotation of the atmosphere with the earth, upon the movement of the superfluous electrons is rather complicated and the exact movements cannot be easily ascertained. The migration of the extra electrons to lower latitudes in variable amounts depending upon variable rates of arrival would account for the movements of the region that have been observed. The principal ionization in the refracting region would be that due to the extra electrons as they moved away from the polar region and while on their journey away from the earth.

When electrons arrive from the sun and strike the earth's magnetic field, they are either deflected away or are captured. The earth's field is in such a direction that an electron striking it over the advancing side of the earth is likely to be deflected away and expelled while if it strikes that part over the following side, it will be deflected in toward the earth and into a denser field and its probability of capture is greater. The greatest number will arrive on the earth not at a point between the magnetic pole and the sun but at a point toward the sunset side. This is substantiated by visual observations on aurora which show definitely a greater number of aurora occurring before midnight than after. The numbers of electrons present in our latitude after sunset would be large because of proximity to the "pole of arrival." The post-midnight depressions in the transmission curves and surfaces on received field strength of the shorter waves would be accounted for by the reduced electron arrival on that side (the early morning side) of the magnetic pole as well as to the greater distance to the "pole of arrival." The number of free electrons reaching the locality of Deal, for instance at this time, the early morning, is not sufficient under average conditions to refract the 16-7/8 meter waves, but at London, Seattle, and Dickinson with the added ionization produced by sunlight, enough are present in June to refract the waves properly during this part of the night.

It is not supposed that all refraction is due to β ray electrons alone. During the day, sunlight will produce large numbers at all altitudes above 16 or 20 miles. The numbers produced will depend upon the strength of the ionizing electromagnetic waves, while their distribution with altitude will be independent of the strength. The variation from minute to minute or day to day will depend upon the variation in strength of the sun's radiation, and so far as is known, that variation is relatively small. The number of electrons produced in the refracting region during the day by the radiation from the sun is likely to be greater than the number usually present from the β ray source, thereby determining to a large extent the refracting region conditions during the day. Under such conditions the height would be decidedly lower during the day than at night. The variations would be less and therefore more uniform refraction would occur. At night, however, the absence of the steady ionization due to sunlight would leave all refraction to the β ray electrons, and the relative variation in ionization from minute to minute becomes more pronounced. The more rapid fading found on those wavelengths that travel well at night than occurs on the shorter waves that travel well in the daytime appears to support this idea.

There are other physical phenomena which fit in well with this hypothesis. β particles from the sun are probably arriving all the time. The occasions when they are brought to our attention are those occasions when extra large numbers arrive, such as during a display of the aurora borealis. They are probably arriving even when the aurora is not seen as Lord Rayleigh[9] reported the existence of the green auroral line in the night sky at all times. The refraction region at night is at the height or some distance above the observed level of the aurora. The aurora is closely associated with earth currents and currents in cables and telegraph lines, while the period of movements observed in this layer correspond closely with the periods of the principal disturbing currents measured in cables. Radio transmission has been observed to be seriously affected at the time of aurora displays. It has also been observed to have a connection with sunspots.

If the views put forward in this paper are correct the refracting region at night should be much lower in the arctic and higher in the tropics than at this latitude (41 deg.). The space charge occurring over the polar region might be forced to an altitude low enough to embrace the absorbing region and produce absorption day and

[9] *Proc.*, Royal Soc. London, Vol. 109, 1925.

night a large part of the time. On occasions of aurora displays of exceptional magnitude, the ionization might be lowered at our latitude as far as the absorbing region and might produce the equivalent of daylight transmitting conditions at night for those waves susceptible to absorption.[10] The height of the refracting or reflecting region should show a connection with sunspot activity. While no direct measurements bearing on this relation are available, it may be significant that Pickard[11] has found a connection between the occurrence of sunspots and long-distance transmission at broadcasting frequencies.

A natural result of our work has been the formation in the mind of a picture of the ionization as it is thought to occur. In one of the earlier efforts two separate and distinct ionized layers were postulated. It was thought they were produced by the two separate ionizing agencies mentioned. An attempt was made to fit this picture to the facts available. It was thought that the difference in absorption noticeable among the short waves could be attributed to the position of the absorbing ionized layer with respect to the absorbing regions for the various short waves. This picture has, however, been discarded in favor of the following.

The greatest number of electrons are pictured as being produced by sunlight, and extending from an altitude of about 16 miles to the limit of the atmosphere. This ionization experiences a diurnal variation due to the rotation of the earth, and a seasonal variation due to the inclination of the earth's axis to the orbit. The intensity at a given height depends largely upon the obliquity of the sun's rays and upon the rate of recombination. The relative intensity of distribution with height is independent of the intensity of sunlight, except as affected in the rarer regions by the rate of recombination. This ionization extends clear across both refracting and absorbing regions. Absorption is produced by this ionization occupying the absorbing region. Within the space occupied by this ionization, a second ionization occurs, irregular in its variation, produced by β particles from the sun. The altitude at which this second ionization occurs is generally considerably above the absorbing region, and only under unusual conditions does it extend down to it. At night the diurnal sunlight ionization disappears and the irregular β ionization remaining is sufficient to refract all but the shortest of usable waves. Rapid fading is the result of rapid variations of intensity in, or position of, this ionization. This ionization is the principal ionization on the dark side of the earth.

[10] Southworth reports observing this phenomenon in April 1926.
[11] I. R. E. Proc., Feb. 1927.

BOOK REVIEW

Radio Manual (U. S. Naval Institute, Annapolis, Md.). 1927 Edition, 141 pages. Price, $5.50.

This Manual is a text book on radio communication first prepared in 1925 by a group of officers in the Department of Electrical Engineering and Physics at the Naval Academy. It is stated in the preface that other texts available were found either too comprehensive or not comprehensive enough for the thirty lessons devoted to the subject at the Academy. Another reason for the writing of this book was the need for an up-to-date work; this requirement has been met by yearly revisions, the 1927 edition being the third.

Of the eighteen chapters, the first three are on wave motion and the characteristics of simple oscillating circuits. In this section, as throughout the Manual, the calculus is used, but merely for purposes of definition. Vector analysis and graphical methods are introduced as required, but the mathematical treatment is very restricted in all the chapters. This is, no doubt, a necessity in such a condensed text. Following chapters on the use of the frequency meter, and apparatus for damped and continuous wave transmission and reception, five chapters are devoted to a discussion of vacuum-tube theory and practice in receivers and transmitters. The remaining chapters are on such subjects as "Radio Telephone Transmitter," "Coil Antennas," etc. The 1926 Report of the Committee on Standardization of the Institute of Radio Engineers is printed as a 21-page supplement, together with membership lists of this Committee for 1923–1925. The book itself has a comprehensive index.

The work is somewhat longer than the number of pages (141) would seem to indicate, the pages being of 8 by 10½ inch size. Even so, a comprehensive treatment of the modern subject of radio communication is impossible in such a space. Morecroft's recently revised text, for example, while not extended deeply into the various specialties, runs to over 1,000 pages. The Manual must accordingly be judged in the light of its purpose: to provide an elementary course in radio theory for student officers, with sufficient practical material to give the men so trained the initial preparation for handling Naval radio equipment later, with the aid of special instruction books and other directions. This object is admirably fulfilled. The treatment is clear throughout, and

the material, carefully arranged, is rendered accessible through numerous and well-worded captions. Each chapter is written and page-numbered as a separate unit, to facilitate frequent revision. If, as a result of the extreme epitomization, statements are occasionally made, as ". . . . it cannot be said that there has been any considerable degree of success attained in the elimination of these effects (atmospheric interference)," which the experienced radio specialist might wish to qualify, this cannot be taken as detracting from the value of such a work in the field of its usefulness. This Manual might, in fact, prove valuable in other elementary radio engineering courses where the allowable time is as limited as at the Naval Academy.

CARL DREHER

Please Refer to Page 14 regarding the discontinuance of publication of these Patent Digests. It is thought that they are not of sufficient general interest to warrant further publication.

DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued November 22, 1927 to December 13, 1927

By

JOHN B. BRADY

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,648,808—WAVE SIGNALING SYSTEM—Louis A. Hazeltine, Hoboken, N. J. Filed Feb. 27, 1925, issued Nov. 8, 1927. Assigned to Hazeltine Corp.

1,649,589—WAVE SIGNALING SYSTEM—Louis A. Hazeltine, Hoboken, N. J. Filed Feb. 27, 1925, issued Nov. 15, 1927. Assigned to Hazeltine Corp.

1,649,810—VACUUM TUBE—E. L. Chaffee, Belmont, Mass. Filed (original) Oct. 14, 1918, renewed Feb. 27, 1925, issued Nov. 22, 1927. Assigned to John Hays Hammond.

1,650,032—RADIO APPARATUS—T. W. New, Cincinnati, Ohio. Filed May 27, 1925, issued Nov. 22, 1927. Assigned to The Teleforce Radio Laboratories Co.

1,650,232—THERMIONIC TUBE—G. W. Pickard, Newton Center, Mass. Filed March 7, 1922, issued Nov. 22, 1927. Assigned to Wireless Specialty Apparatus Co.

1,650,353—WAVE SIGNALING SYSTEM—L. A. Hazeltine, Hoboken, N. J. Filed Feb. 27, 1925, issued Nov. 22, 1927. Assigned to Hazeltine Corp.

1,650,250—TWO WAY CIRCUIT ARRANGEMENT FOR WIRELESS TELEPHONY—O. Von Bronk, Berlin, Germany. Filed Feb. 20, 1923, issued Nov. 22, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie.

1,650,395—FIXED CAPACITY CONDENSER AND METHOD OF MAKING THE SAME—S. Siegel, New York, N. Y. Filed April 14, 1925, issued Nov. 22, 1927. Assigned to Aerovox Wireless Corp.

1,650,461—ANTENNA DEVICE—A. R. Nilson, Brooklyn, N. Y. Filed Oct. 10, 1925, issued Nov. 2, 1927.

1,650,465—VARIABLE CONDENSER—W. Reese, Brackenridge, Pa. Filed Dec. 27, 1926, issued Nov. 22, 1927.

1,650,701—RADIO SIGNALING SYSTEM—J. F. Farrington, Flushing, N. Y. Filed April 22, 1925, issued Nov. 29, 1927. Assigned to Western Electric Co., Inc.

1,650,754—VACUUM TUBE UNIT—A. A. Kent, Ardmore, Pa. Filed Apr. 20, 1923, issued Nov. 29, 1927.

1,650,898—TUNED RADIO FREQUENCY CIRCUITS—D. R. Lovejoy, New York, N. Y. Filed June 4, 1925, issued Nov. 29, 1927. Assigned to Lovejoy Development Corp.

1,650,921—VACUUM TUBE—A. Winkelmann, Hoboken, N. J. Filed May 9, 1923, issued Nov. 29, 1927.

1,650,862—SPIRAL PLATE CONDENSER—F. A. Borsych, Minatare, Neb. Filed Feb. 15, 1926, issued Nov. 29, 1927.

1,650,925—COMBINED TELEPHONE AND WIRELESS RECEIVER—C. H. Allen, Pittsburgh, Pa. Filed Mar. 10, 1923, issued Nov. 29, 1927.

1,650,934—SYSTEM OF MODULATION—L. W. Chubb, Swissvale, Pa. Filed Aug. 30, 1923, issued Nov. 29, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,650,944—APPARATUS FOR RADIO TRANSMISSION—Abd-El-Rahman Z. A. Khalifah, Philadelphia, Pa. Filed Nov. 29, 1922, issued Nov. 29, 1927.

1,650,983—INSULATING STRUCTURE FOR HIGH POTENTIAL CONDENSER TERMINALS AND THE LIKE—W. Dubilier, New York, N. Y. Filed Nov. 18, 1921, issued Nov. 29, 1927. Assigned to Dubilier Condenser Corp.

1,651,065—METHOD OF EXHAUSTING THERMIONIC TUBES AND LIKE DEVICES—H. E. Metcalf, Oakland, Calif. Filed Jan. 21, 1927, issued Nov. 29, 1927. Assigned to Magnavox Co.

1,651,308—AUDIO AMPLIFIER—L. Winkelmann, Hoboken, N. J. Filed Apr. 10, 1922, issued Nov. 29, 1927.

Re-16,805—RADIO TELEGRAPH AND TELEPHONE RECEIVING SYSTEM—F. G. Simpson, Seattle, Wash. Filed Nov. 4, 1922, issued Nov. 29, 1927. Assigned to Simpson Radio Corp.

1,651,810—AMPLIFYING SYSTEM—Alfred Crossley, of Washington, D. C. Filed Nov. 10, 1925, issued Dec. 6, 1927. Assigned to Wired Radio, Inc.

1,651,953—SUPPORT FOR RADIO APPARATUS—A. C. Hayden, of Brockton, Mass. Filed Dec. 13, 1926, issued Dec. 6, 1927.

1,651,975—VARIABLE CONDENSER—H. M. Specht, of Pelham, N. Y. Filed Dec. 19, 1923, issued Dec. 6, 1927.

1,652,118—RADIO CONDENSER—M. Guett, of Hartford, Connecticut. Filed July 1, 1925, issued Dec. 6, 1927. Assigned to Hart & Hegeman Manufacturing Co.

1,652,155—ELECTRICAL SYSTEM AND SIGNALING METHOD—A. F. Van Dyck, of Rye, New York, Filed May 27, 1924, issued Dec. 6, 1927. Assigned to Radio Corporation of America.

1,652,158—VARIABLE ELECTROSTATIC CONDENSER—W. Aull, Jr., New York, N. Y. Filed Nov. 29, 1922, issued Dec. 13, 1927. Assigned to Dubilier Condenser Corp.

1,652,164—METALLIC OSCILLION AND METHOD OF CONSTRUCTING SAME.—H. S. Coyer, New York. Filed Dec. 18, 1920, issued Dec. 13, 1927.

1,652,212—ELECTRICAL CONDENSER AND METHOD OF MAKING IT—W. H. Priess, Belmont, Mass. Filed Mar. 30, 1921, issued Dec. 13, 1927. Assigned to Wireless Specialty Apparatus Co.

1,652,219—RADIO RECEIVING APPARATUS—A. A. Thomas, New York, N. Y. Filed May 20, 1922, issued Dec. 13, 1927.

1,652,257—COMPOSITE RADIO PANEL AND SOUND MODIFIER—W. B. Stevenson, Philadelphia, Pa. Filed Aug. 20, 1925, issued Dec. 13, 1927. Assigned to Victor Talking Machine Co.

1,652,376—VARIABLE CONDENSER—J. D. Sartanoff, New York, N. Y. Filed Sept. 29, 1925, issued Dec 13, .1927. Assigned one-half to Khetah Corp.

1,652,388—METHOD AND ARRANGEMENT FOR THE MULTIPLE AND DIRECTED RECEIVING OF RADIO SIGNALS—V. I. Bashenoff, Moscow, Russia. Filed June 18, 1923, issued Dec. 13, 1927.

1,652,497—THERMIONIC DEVICE—H. M. Ryder, Wilkinsburg, Pa. Filed June 25, 1921, issued Dec. 13, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,652,516—RADIO TRANSMITTING SYSTEM—F. Conrad, Pittsburgh, Pa. Filed Dec. 23, Dec. 13, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,652,863—ELECTRIC CONDENSER—F. J. Kent, Verona, N. J. Filed Mar. 31, 1923, issued Dec. 13, 1927. Assigned to Unitrol Condenser Corp.

1,652,880—VACUUM TUBE AND TUBE CIRCUITS—A. G. Thomas, Lynchburg, Va. Filed July 25, 1923 issued Dec. 13, 1927.

1,652,901—VARIABLE CONDENSER—R. H. Langley, Schenectady, N. Y. Filed Sept. 26, 1923, issued Dec. 13, 1927. Assigned to General Electric Co.

1,652,927—CRYSTAL DETECTOR AND CLIP—J. Buchanan, North Braddock, Pa. Filed June 2, 1924, issued Dec. 13, 1927.

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED
December 7, 1927

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California	Oakland, Rola Co., 4250 Hollis Street	Engholm, Bernard A.
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Louisiana	New Orleans, U. S. Customhouse, Room 217	Deiler, Theodore G.
	New Orleans, 2222 Sapeyrouse Street	Gallo, Louis J.
New Jersey	Maywood, [illegible]6 Maywood Avenue	Stantley, Joseph J.
Ohio	Cleveland, 3041 Fairfax Road	Victoreen, John A.

Elected to the Member grade

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	Mayfield, 285 College Avenue	Paul, Ernest Edward
	Oakland, 2552 Foothill Boulevard	Sodeberg, Arnold J.
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	San Francisco, 480 Second Street	MacGregor, Robert
	Santa Barbara, 1010 Bath Street	Kempton, Walter D.
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	Denver, c/o Denver Tramway Corp.	Love, Nelson R.
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	New London, U. S. S. Downes	Dillow, Arthur P.
	Stamford, 20 Worth Street	Stephenson, Martin L.
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Georgia	Royston	Berryman, Forrest G.
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	Chicago, 140 S. Kilbourn Avenue	Herrick, R. H.
	Chicago, 3829 Fulton Street	Keogh, Raymond J.
	Chicago, 1942 N. Lowell Avenue	Porter, G. O.
	Chicago, 5531 Washington Boulevard	Stevens, Fred B.
	Zion, 3010 Emmaus Avenue	Mason, C. T.
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Louisiana	New Orleans, 119 So. St. Patrick Street	Jensen, Valdemar
Maryland	Baltimore, 2502 Dulaney Street	Seitz, J. Fred
	Cumberland, Box 345	Craig, Joseph B.
	Westminster, 93 Liberty Street	Utz, William C.
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	Winchester, 407 Main Street	Burgoyne, Roger M.
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	Detroit, 5316 Seebaldt Avenue	Puddiford, Stanley W.
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	Lakefield	Tkach, George
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	St. Louis, 3827 Vest Avenue	Henrie, Edward F.
	St. Louis, 5950 Kingsbury Avenue	Whitlock, Fred
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	Bronx, 740 Prospect Avenue	Frimmerman, Frank
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	Brooklyn, 5818 13th Avenue	De Martino, Michael W.
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	Brooklyn, 359 Ovington Avenue	**Lussier, Elphege A.**
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	Buffalo, 522 Northampton Street	**Peters, Albert H.**
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	Buffalo, 167 Taunton Place	**Wagener, W. Hall**
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	Mamaroneck, 124 Barry Avenue	**Annucci, Louis M.**
	Middletown, 25½ Wisner Avenue	**Clark, Ogden V.**
	New York City, Radio Institute of America, 326 Broadway	**Black, O. M.**
	New York City, 3852 Tenth Avenue	**Di Giacomo, Aniello**
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	New York City, North Brother Island	**Halpin, William A.**
	New York City, A. T. & T. Co., 195 Broadway	**Kelley, Leo A.**
	New York City, 11 E. 199th Street	**Potts, John H.**
	New York City, 290 East 2nd Street	**Robins, J. Francis**
	New York City, 2671 Broadway	**Spiegelthal, Edwin**
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	Rochester, 133 Wilder Terrace	**Cass, Lewis S.**
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	Williamsville, 5565 Main Street	**Hopkins, Nelson S.**
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Ohio	Akron, 20 N. Balch Street	**Davis, Kenneth**
	Akron, 145 Center Street	**Schoenduve, George C**
	Akron, Lake States Gen. Electric Sup. Co.	**Schoenduve, H. W.**
	Cinti, 2175 Central Avenue	**Wanstrath, Wm. T.**
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Australia	Hobart Tasmania, Box 401	**Hooper, Edgar M.**
	Victoria, Braybrook, Braybrook Radio Station	**Brooker, Vivian M.**
	Victoria, East Malvern, 8 Viva Street	**Endean, Douglas A.**
Canada	British Columbia, Vancouver, Canadian Natl. Ry.	**Claringbull, D.**
	Ontario, Toronto, 727 Indian Road	**Kynnerseley, Charles**
	British Colombia, Ocean Falls, Box 171	**Marriott, John S.**
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Czechoslovakia	Prague, Ministry of Posts and Telegraphs	**Strnad, Joseph**
England	Bucks, Wolverton, New Bradwell, 30 King Edward St	**Clarke, D. O.**
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Ireland	Co. Tipperary, Main Street, Emly	Gargan, Edward F.
Japan	Tokyo, 500 Oimachi	Yoshio, Imaoka
Mexico	Mexico D. F., Calle del Buen Tono No. 12.	Buchanan, J. C.
New Zealand	Wellington, Radio Dept., Union Steamship Co.	Matthews, P. H.

Elected to the Junior grade

Georgia	Atlanta, 1593 Rodger Avenue, S. W	Matthews, Allen A., Jr.
Illinois	Chicago, 5783 Ridge Avenue	Riddel, O. A.
Kansas	Lawrence, 1622 New Hampshire Street	Douglas, Nowel
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New York	Canandaigua, Canandaigua Radio Service.	Thompson, John M.
Ohio	Fort Clinton, 120 Maple Street	Anderson, G. J. C.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 16 February, 1928 Number 2

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A. Hoyt Taylor

Recipient of the Morris Liebmann Memorial Prize, 1927.

A. Hoyt Taylor

RECIPIENT OF THE MORRIS LIEBMANN MEMORIAL PRIZE, 1927

A. Hoyt Taylor was born January 1, 1879 in Chicago, Illinois. He attended grammar school in Wilmette, Illinois, high school in Evanston, Illinois and graduated with the B. S. degree from Northwestern University. In 1900 he accepted a position as instructor at Michigan State College, teaching physics and electrical engineering. In 1903 he became instructor in electrical engineering at the University of Wisconsin and for some time specialized in electrical measurements with particular attention to alternating currents, and published one or two papers on high-frequency measurements, besides various publications on precision measurements with alternating currents.

In 1908 while assistant professor at the University of Wisconsin, he was granted a year's leave of absence to study in Germany, where he received the Ph. D. degree at the University of Gottingen, specializing in applied electricity, mathematics, and mathematical physics. Upon returning to the United States in the fall of 1909 he was appointed professor of physics and head of department at the State University of North Dakota.

Early in 1911 Dr. Taylor started his radio work at the University of North Dakota and constructed the station which was later known as 9XN. From the beginning of this work, particuar attention was given to the study of wave propagation phenomena, fading, influence of weather conditions, studies of directional systems, etc. Various publications covering this work, prior to 1917, appeared in *The Electrical World*, *The Physical Review*, *Journal* of the University of North Dakota, and in the PROCEEDINGS of the Institute of Radio Engineers.

Dr. Taylor entered the Naval Reserve as Lieutenant in March of 1917, and was district communication officer at Great Lakes Naval Training Station until he was transferred to the east coast with headquarters at Belmar, New Jersey as transatlantic communication officer in the fall of 1917.

In the summer of 1918 he was promoted to Lieutenant Commander and sent to the Naval Operating Base at Hampton Roads, Virginia. While at Hampton Roads he acted as head of the Experimental Division of the Naval Air Station studying, particularly, aircraft radio development work.

Early in 1919 he was ordered to Washington, D. C. and was placed in charge of an aircraft radio laboratory with additional duties in a consulting capacity for other radio activities of the naval service. During this period he was promoted to the rank of Commander in the reserve force, remaining in active duty in this rank until July of 1922. Since then he has remained a reserve officer on inactive duty.

Dr. Taylor was made Superintendent of the Radio Division at the Naval Research Laboratory, Bellevue, Anacostia, D. C.

During the past few years his most important work has been the long series of experiments which led to the publication of papers dealing with the theory of short wave transmission. For his work in this connection the Board of Direction of the Institute awarded Dr. Taylor the Liebmann Memorial Prize for 1927.

Dr. Taylor has been a frequent contributor to the PROCEEDINGS of the Institute, and is a Fellow in the Institute.

CONTRIBUTORS TO THIS ISSUE

Aiken, Charles B.: Born at New Orleans, Louisiana, October 6, 1902. Marine radio operator, summers 1918–1921. Received B. S. degree in physics, Tulane University, 1923; M. S. degree, Harvard, 1924; M. A. degree, Harvard, 1925. Winner of Harvard Engineering School Prize Scholarship for 1924–1925. Whiting fellow in physics, 1924–1926. Research Engineer, Mason, Slichter and Hay, Consulting Engineers, 1926 to date. Associate member of the Institute.

Austin, L. W.: Born at Orwell, Vermont, October 30, 1867. Received A.B. degree, Middlebury College, 1889; Ph. D. degree, University of Strassburg, 1893. Instructor and assistant professor, University of Wisconsin, 1893–1901; research work, University of Berlin, 1901–1902; Bureau of Standards, Washington, D. C. since 1904; head of U. S. Naval Radio Research Laboratory, 1908–1923; chief of Radio Physics Laboratory, 1923 to date. Dr. Austin was President of the Institute in 1914 and served on its Board of Direction from 1915 to 1917. In 1927 he was awarded the Institute Medal of Honor. His contributions to the PROCEEDINGS have been frequent. He is a Fellow of the Institute.

Horton, J. H.: Born at Ipswich, Massachusetts, December 18, 1889. Received B.S. degree in electro-chemistry, Massachusetts Institute of Technology, 1914. Instructor, Massachusetts Institute of Technology, 1914–1916; Research engineer with Western Electric Company, 1916–1925 (on leave of absence 1917–1918 as technical expert with U. S. Navy working on methods for detection and location of submarines at experimental stations in this country). At present research engineer with Bell Telephone Laboratories, engaged in development of systems for the transmission of pictures and television; also in development of methods for the precision measurement of frequency. Associate member of the Institute.

Hulburt, E. O.: Born October 12, 1890. Received Ph. D. degree in physics, Johns Hopkins University, 1915. Taught undergraduate and graduate courses at Western Reserve University, Johns Hopkins University and University of Iowa. With American Expeditionary Force in France for two years as Lieutenant and later Captain in the Signal Corps. At present Dr. Hulburt is Superintendent of the Heat and Light Division, Naval Research Laboratory, Bellevue, D. C. He is the author of a number of papers on experimental and theoretical work in spectroscopy, physical optics and radio telegraphy.

McIlwain, Knox: Received B.S. degree, Princeton University, 1918; B.S. in E. E., University of Pennsylvania, 1921. U. S. Navy 1917–1919. Engineering department of Bell Telephone Company of Pennsylvania, 1921–1924. Instructor at Moore School of Electrical Engineering, University of Pennsylvania, 1924 to date.

Maris, H. B.: Born in 1885. Received A.B. degree, University of Michigan, 1909; M.S. degree, 1910; Ph.D. degree, Johns Hopkins University, 1927. Associate Professor of Physics at Birminghan-Southern College, 1922. Professor of physics, Emory and Henry College, 1923. Consulting physicist, Naval Research Laboratory, 1925 to date; researches, photo-elastic studies, and theoretical study of the upper atmosphere.

Marrison, W. A.: Served in Royal Flying Corps, later Royal Air Force, in Canada, 1917–1918; received B.S. degree in physics, Queens University, Canada, 1920; A.M. degree in physics and mathematics, Harvard University, 1921. Research engineer with Western Electric Company 1921–1925; research engineer with Bell Telephone Laboratories, engaged in the study of picture transmission and methods for the production of constant frequency, 1925 to date.

Nyman, Alexander: Born in Finland, December 29, 1893. Received B.S. degree, Manchester University, England, 1915. With Westinghouse Electric and Manufacturing Company, Pittsburgh, 1915–1923, with exception of brief period of service in U. S. Army. Associated with the Dubilier Condenser Corporation, 1923 to date, in capacity of technical manager, and later consulting engineer. Manager of Radio Patents Corporation, 1925 to date. Mr. Nyman is a member of the Institute.

Rodwin, George: Born at New York City, September 4, 1903. Received A.B. degree, Columbia University, 1923; E. E. degree, 1925. With Radio Corporation of America doing research work in connection with fading recorder equipment, standard receiver testing methods, receiver development, and general broadcast station equipment, 1925 to date. Associate member of the Institute.

Smith, Theodore A.: Born at Brooklyn, New York, February 17, 1905. Received M.E. degree, Stevens Institute of Technology, 1925. With Radio Corporation of America, Technical and Test Department, engaged in work on fading recording, short wave reception, field intensity measurements, and miscellaneous work on broadcast station engineering, 1925 to date. Associate member of the Institute.

Thompson, Walter S., Jr.: Received E. E. degree, Lehigh University, 1924. With Engineering Department, Bell Telephone Company of Pennsylvania, 1924–1925. Instructor, Moore School of Electrical Engineering, University of Pennsylvania, 1925.

von Ardenne, Manfred: Born at Hamburg, Germany, January 20, 1907. Educated at University of Berlin. Engaged in radio engineering since 1921; own research laboratory, 1923 to date, specializing in audio-frequency amplification and reproduction. Associate member of the Institute.

Wymore, Ivy Jane: Born in Mahaska County, Iowa. Received B.S. degree, Drake University, 1918; M.S. degree, George Washington University, 1925. With Division of Metallurgy, Bureau of Standards, 1919–1924; Laboratory for Special Radio Transmission Research, 1924 to date.

INSTITUTE ACTIVITIES

January Meeting of the Board of Direction

At the meeting of the Board of Direction of the Institute held on January 4, 1928 the following were present: Ralph Bown, President, A. N. Goldsmith, Secretary, R. A. Heising, R. H. Marriott, L. E. Whittemore, A. H. Grebe and J. M. Clayton, Assistant Secretary.

The ballots for election of 1928 officers and managers of the Institute were counted, with the following results: President, A. N. Goldsmith; Vice President, L. E. Whittemore; Members of Board of Direction to serve until January 1, 1931: J. H. Dellinger and R. H. Manson.

To fill the unexpired term caused by the death of Colonel John F. Dillon, the Board appointed J. V. L. Hogan. The board appointed the following as managers with one year terms: Arthur Batcheller, W. G. Cady, A. H. Grebe and L. A. Hazeltine.

Commending the services of President Bown, the Board passed the following resolution:

"The Board of Direction expresses its earnest appreciation of the competent and sympathetic direction of its activities during the time Dr. Ralph Bown was President of the Institute."

The following were transferred or elected to the higher grades of membership in the Institute: transferred to the grade of Fellow: Joseph D. R. Freed. Transferred to the grade of Member: F. W. Cunningham and A. M. Patience. Elected to the grade of Member: E. K. Lippincott, G. Schottel and S. Siegel.

One hundred and twelve associate members and twelve junior members were elected.

1928 Convention

On January 9th, 10th, and 11th the Third Annual Convention of the Institute was held in New York. The program opened with an address by the retiring president, Dr. Ralph Bown, followed by the presentation of the 1927 Morris Liebmann Prize to Dr. A. Hoyt Taylor. Dr. Bown next introduced President A. N. Goldsmith who, in turn, introduced Vice-President L. E. Whittemore. W. D. Terrell, of the Department of Commerce, read a paper on "The International Radiotelegraph Conference of Washington, 1927."

In the afternoon of January 9th a tour of inspection to the Bell Telephone Laboratories Experimental Station Group and the

National Broadcasting Station WJZ at Bound Brook was taken by over three hundred and fifty delegates to the convention.

In the evening papers by Dr. J. H. Dellinger, Dr. F. K. Vreeland, and E. H. Loftin and S. Y. White were read.

An inspection trip to Roxy's Theatre with a technical session on the making of talking moving pictures took place on the morning of January 10th. In the afternoon a trip was taken to the Talking Moving Picture Studio of the Radio Corporation of America and to the plant of the Polymet Manufacturing Company. That evening Captain Richard H. Ranger presented a paper with demonstrations on the transmission of photographs by radio.

The delegates, on Wednesday morning, inspected the studios of The National Broadcasting Company, the plant of F. A. D. Andrea and the plant of the Aerovox Wireless Corporation, luncheon being served by the Aerovox Corporation. In the afternoon a symposium relating largely to inter-electrode capacities in in tubes was presented, the following delivering papers: Lincoln Walsh, Harold A. Wheeler, E. T. Hoch, and J. C. Warner. Following the afternoon sessions a trip to the Paramount News Bureau was arranged.

On the final evening dinner was held at the Hotel Roosevelt. George C. Furness presided at this meeting. The program included short introductions of prominent Institute officials, an address by Dr. A. N. Goldsmith, entertainment by prominent radio broadcast stars, and music by the Vincent Lopez Orchestra. Following the dinner dancing was provided.

The total registration at the convention was over seven hundred. Over four hundred and fifty members of the Institute and their guests attended the dinner.

Convention Papers Available

The following papers presented at sessions of the convention are available free of charge to members of the Institute:

"The International Radiotelegraph Conference of Washington, 1927," by W. D. Terrell.

"On the Distortionless Reception of a Modulated Wave and Its Relation to Selectivity," by Frederick K. Vreeland.

"Some Characteristics and Applications of Four-Electrode Tubes," by J. C. Warner.

"Measurement of Vacuum-Tube Capacities by a Transformer Balance," by Harold A. Wheeler.

"A Direct Capacity Bridge for Vacuum-Tube Measurements," by Lincoln Walsh.

"Direct Coupled Detector and Amplifiers with Automatic Grid Bias," by Edward H. Loftin and S. Young White.

Upon application to the offices of the Institute a copy of any of the above papers in pamphlet form will be mailed to members.

Meeting of Section Representatives

One meeting of the Institute was devoted to a conference of representatives of the sections of the Institute. Addresses were delivered by President Goldsmith, Past Presidents Ralph Bown and Donald McNicol, and by a number of the chairmen of sections.

The conference extended over a period of four hours during which time a vast amount of information relative to section operation and management, together with an account of the problems of individual sections, was summarized.

A summary of the activities of this conference will be available in the near future for members interested in the formation of a section.

It is planned that this very important feature of the Institute conventions will be held at future annual conventions.

Section Meetings

Atlanta Section

R. M. Wise, chief engineer of the E. T. Cunningham Company, delivered a paper on "Shield Grid Tubes, AC Tubes and Oxide Filament Rectifier Tubes" at the meeting of the Atlanta Section, held on January 3rd in the Chamber of Commerce Building, Atlanta, Georgia.

Fourteen members of the Institute and twenty-two guests attended this meeting.

Buffalo-Niagara Section

On December 14th a meeting of the Buffalo-Niagara Section was held in Foster Hall, University of Buffalo. L. C. F. Horle, chairman, presided.

W. R. Jones, research engineer of the Federal Radio Corporation presented a paper on "Notes on Design and Production of Uni-Control Broadcast Receivers."

Messrs. Horle, Henderson, Porter and others participated in the discussion which followed the presentation of the paper.

Forty-five members of the Section attended this meeting.

A meeting of the Section was held on January 18th in Foster Hall, University of Buffalo, at which time Dr. Leo Dana, chief physicist of the Linde Air Products Company read a paper entitled, "Application of Rare Gases to Radio."

Messrs. Horle, Lidbury, Porter, Hector and others discussed the paper.

Eighty-five members and guests attended.

The next meeting of the Section will be held on February 15th in the University of Buffalo. Carl Dreher, staff engineer of The National Broadcasting Company, will present a paper on "As the Broadcaster Sees It."

Canadian Section

The Canadian Section held a meeting on December 7th at which V. G. Smith delivered one of the junior lectures on "Series and Parallel Resonance." F. K. Dalton presented a paper entitled, "Marconi Beam Stations."

A. M. Patience, Chairman of the Section, presided.

In the discussion following the two papers, the following members took part: D. Hepburn, C. I. Soucy, V. G. Smith, C. C. Meredith and others.

The attendance at this meeting was fifty-six.

Chicago Section

On December 16th a meeting of the Chicago Section was held in the Auditorium of the Western Society of Engineers. Professor G. M. Wilcox presided.

Dr. Frederick W. Kranz, of Riverbank Laboratories, presented a paper on, "Some Characteristics of Speech and Hearing."

Following the discussion the annual election of officers was held, the result being that John H. Miller was elected Chairman, Harold L. Olesen, Vice-Chairman; H. E. Kranz, Secretary-Treasurer; and the Executive Committee with its membership as follows was appointed: G. M. Wilcox and E. L. Koch.

Cleveland Section

A meeting of the Cleveland Section was held on January 6th in the Ohio Bell Telephone Building. John R. Martin presided.

D. A. Leach, equipment engineer of the Ohio Bell Telephone Company, delivered a talk entitled "Automatic or Machine Switching." The talk included lantern slides, followed by an inspection of the new automatic equipment in the building.

Preceding the technical meeting, an informal dinner was held in the Hotel Winton.

Sixty-three members of the Section attended the meeting.

DETROIT SECTION

Dr. N. H. Williams presented a paper, "Some Characteristics of the Screen Grid Tube" at a meeting of the Detroit Section held on December 16th in the dining room of the Michigan Bell Telephone Company Building, Detroit.

Thomas E. Clark, Chairman of the Section, presided.

Over one hundred members of the Section and their guests attended the meeting.

Preceding the meeting a dinner, at which seventy-five persons were present, was held.

LOS ANGELES SECTION

On November 21st a meeting of the Los Angeles Section was held in the Elite Cafe, 633 South Flower Street, Los Angeles.

D. C. Wallace, Vice-Chairman, presided.

Dr. Leonard F. Fuller delivered a paper entitled "Vacuum Tubes and Their Application to the Power Field."

L. Elden Smith delivered a paper describing the short wave work accomplished on the Yacht *Ripple* through the South Seas, and also a description of radio beacons at Wheeling Field, Hawaiian Islands.

Seventy-two members of the Section were present.

Committee Work

I. R. E. SUBCOMMITTEE ON RECEIVING SETS

A meeting of the Subcommittee on Receiving Sets was held at the Institute Office on January 11th. Those present were: J. H. Dellinger (Chairman), E. E. Hiler (Secretary), E. Austin, I. G. Maloff, W. D. Kirschbaum, W. A. Diehl, C. A. Wright, A. H. Lynch, L.C.F.Horle, George Crom, H.B.Coxhead, and L.M.Hull.

The Subcommittee modified Section D of the printed May 20, 1927 preliminary report, by the adoption of an alternative method of measuring input field intensity, viz, the introduction of input voltage by a coupling resistor in the output circuit of the radio-frequency source. This was added because some laboratories have found this method to be convenient and to give results in agreement with those obtained by the use of a coupling coil.

The preparation of a section on Test Procedures was begun. This is a difficult undertaking and will require considerable more

work. It is expected that procedures will be worked out in sufficient detail to be applicable to the several types of receiving sets. The Committee is giving attention to correlation of its recommendations with those of the Subcommittee on Vacuum Tubes.

Extensions were made in the Bibliography. It has been found in listing references on receiving set testing that articles on other receiving apparatus must necessarily be included. This led to a recommendation that the Standardization Committee have a special subcommittee on Bibliography, in order that such work may be correlated for the whole field to which the Committee is giving attention.

I. R. E. Subcommittee on Vacuum Tubes

On December 6th a meeting of the Subcommittee on Vacuum Tubes, L. A. Hazeltine, Chairman, was held in the offices of the Institute. All members of the Committee were present or were represented.

All suggestions which had been received, for modification of the preliminary draft of May 20, 1927 were considered and some were adopted.

The connections of Fig. 4, a bridge method for measuring grid-plate capacity, were modified by interchanging the grid and plate and by substituting an adjustable resistance in series with the standard capacity for the adjustable capacity used for phase balance. This is the arrangement originally proposed by Lincoln Walsh, which was presented by him in the symposium on "Vacuum-Tube Capacity measurements" at the Convention. The main dimensions of a shielding plate, in which the vacuum tube is to be mounted, were specified, these being in accordance with the drawing appearing in the current "Nema Handbook of Radio Standards."

The subject of power rating of vacuum tubes used particularly to supply loudspeakers was discussed at length. It was decided to give, in a single section, specifications for "Maximum Undistorted Power Output" and "Conventional Power Output," the former to be essentially those given in the preliminary draft (Section 18), the latter to correspond with those for "Normal Output" in the report of the Subcommittee on Receiving Sets. The difference lies, essentially, in the value of resistance to be put in the external plate circuit: maximum undistorted power output calls for twice the plate resistance, or a higher value, if specified by the manufacturer, while conventional power output is a lower

value taken with an external resistance equal to the plate resistance, this giving the greatest output for a given input alternating voltage. In order to further bring the work of the two subcommittees into accord, Mr. Engel was asked to cooperate with Mr. Van Dyck on the Subcommittee on Receiving Sets in studying the desirability of directly measuring harmonics produced in the plate circuit, in place of inferring their magnitude by the change in the direct component of plate current, as specified in the preliminary draft.

It was voted unanimously to recommend to the main Committee on Standardization the use of the word "Capacitance" in place of "Capacity." This is in accord with the usage of the A. I. E. E. and of the N. E. M. A.

It was decided to define "screen-electrode vacuum tube," this being the name chosen for the new four-electrode tube in which the fourth electrode, or "screen," serves to screen the grid electrostatically from the plate.

Personal Mention

Ralph R. Batcher, formerly engineer with the A. H. Grebe Company, is now vice-president of the Decatur Manufacturing Company, Inc., of Brooklyn, New York.

B. R. Hubbard is now director of laboratory for the Submarine Signal Corporation of Boston, having returned from leave of absence at Massachusetts Institute of Technology, where he was an assistant instructor.

R. A. Hackbusch, formerly on the staff of the Canadian Westinghouse Company, Ltd., is now with Canadian Brandes, Ltd., of Toronto.

Charles C. Henry has resigned as radio sales engineer of the Sonora Phonograph Company, Inc. of Saginaw, Michigan, and has joined the staff or Grigsby-Grunow-Hinds Company of Chicago, in the same capacity.

William H. Fortington, late director of research of the Operadio Corporation, of Chicago, is now practicing as a consulting engineer in Chicago.

Lieutenant Leonard H. Bourchier, U. S. Marine Corps, has been transferred from the Radio Station, Belize, Honduras, to Radio Station, 2nd Brigade, U. S. Marines, at Managua, Nicaragua.

A. I. E. E. 1928 Winter Convention

With headquarters in the Engineering Societies Building, 33 West 39th Street, New York, the A. I. E. E. Winter Convention will be held February 13–17.

The "Communications" portion of the technical sessions will take place on Thursday, February 16th and will include two papers on "Transatlantic Telephony," followed by an exchange of greetings over the New York-London radiotelephone circuit between the President of the American Institute of Electrical Engineers and the President of the British Institution of Electrical Engineers. Arrangements have been made for these and other exchanges of greetings to be heard by those present at the sessions of the A. I. E. E. in New York, and also by members of the Institution of Electrical Engineers who will have a regular meeting simultaneous with the New York meeting.

A paper by C. R. Hanna on "A New Horn Type of Loudspeaker" will be presented, followed by a paper by H. B. Nyquist on "Certain Topics in Telegraph Transmission Theory."

A PRECISION METHOD FOR THE MEASUREMENT OF HIGH FREQUENCIES

By

Charles Bayne Aiken

Summary—*A precision method for the measurement of the frequency of an oscillating circuit is discussed. The theory on which the method is based is discussed and there is developed an equation which relates the frequency of the beat note between two oscillators to the natural frequency of a circuit which is loosely coupled to one of them. This equation is considered in some detail and certain of its properties are deduced. Curves are drawn for three typical cases. The cause and avoidance of certain errors are considered. The method is extended to the case of a non-oscillating circuit. Finally there is suggested a method for the measurement of small values of mutual inductance.*

FOR a number of years there has been employed in the Cruft Laboratory of Harvard University a zero beat method of frequency measurement which is susceptible of great precision. A brief mention of this method has been made by Professor G. W. Pierce in his paper, "Piezo-Electric Crystal Resonators and Crystal Oscillators Applied to the Precision Calibration of Wave Meters."[1] In the present paper a more detailed consideration of the method is given, together with a development of the theory involved. Suggestions are made for the application of the results of this development to the measurement of very small values of mutual inductance.

Part 1—Description of the Method

There will first be described the measurement of the frequency of a source of sustained oscillations. We shall, as a matter of custom, speak of the meter which is to be used as a wavemeter, but shall call the quantity under measurement a frequency since the equations have been developed in terms of 2π times the frequency.

A vacuum-tube generator will oscillate at such a frequency as to make the total effective reactance of the oscillating circuit zero. This reactance is determined not only by the constants of the oscillating circuit proper, but also by the constants of the tube and of whatever circuits may be coupled to the oscillating system, including the plate circuit. In the present method a wavemeter is coupled to the oscillating circuit and advantage is taken of the effect which the tuning of this added circuit has upon the frequency of the

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[1] *Proc. Amer. Acad.*, 59–4, 1923.

oscillator. It is well-known that resonance effects in coupled circuits depend upon the constants of both circuits and that if the coupling is sufficiently loose the current in one circuit will follow a single peaked resonance curve when the other circuit is tuned. It is readily shown that this peak occurs when both circuits have the same period as either would have if the coupling were decreased to zero and that the total effective reactance of each circuit is equal to zero for this condition. If now one of the circuits is thrown out of resonance its effective reactance will no longer be zero nor will that of the other circuit. This is indicated by the fact that the currents in both circuits will decrease. If now the circuit which is arbitrarily thrown out of resonance is a wavemeter circuit and the other is the oscillating circuit of a vacuum-tube generator it follows that the frequency of oscillation will change in such a way as to keep the effective reactance of the oscillating circuit zero. In the neighborhood of mutual resonance this effect of the wavemeter tuning on the oscillating frequency may be large, but when the wavemeter is tuned to a frequency far removed from that which the oscillator would have alone, the effect of the wavemeter in altering the frequency will become vanishingly small. It is at once apparent that this phenomena furnishes us with a method for determining the frequency of the oscillator. This determination is made as follows:

With the wavemeter removed, or open circuited, zero beats are obtained between the fundamental or any convenient harmonic of the oscillator and some other source of high-frequency oscillations. This second source is to be used merely as a reference frequency which will be serviceable in making evident the variations in the frequency of the main oscillator, and it is assumed that the reaction of the oscillator on the auxiliary source is negligible. Instead of using another oscillator a distant transmitting station may be used and the oscillator tuned to zero beats with the signal brought in by a receiving set. After the zero beat adjustment has been made the wavemeter is loosely coupled to the oscillator and the tuning of the meter changed slowly. As resonance is approached the beat frequency will depart from zero and gradually rise to a maximum. As the tuning is continued, the beat frequency drops sharply off and passes through zero and then rises rapidly to another maximum after which it falls off gradually to zero. When the middle silent point occurs the oscillator and the wavemeter are both tuned to the frequency of the auxiliary signal, providing the fundamentals of both oscillations have been employed. If the wavemeter has

already been calibrated the frequency of the oscillator is determined as well as that of the auxiliary signal. If the last mentioned is a standard frequency broadcast then the oscillator frequency i- again determined and a point is obtained for calibrating the waves meter. By employing various harmonic ratios between the oscillator and the incoming signal several points can be obtained for the wavemeter calibration. If the harmonics of the incoming signal are too weak another oscillator may be tuned to zero beat with this signal, and the various harmonics of this oscillator made to beat with those of the first oscillator. The beat note is heard by inserting a pair of telephones in the output circuit of one or the other oscillators.

If the beat is adjusted to a value which is below the audible range of frequencies but is not at zero, the two maxima in the beat frequency which are obtained when the wavemeter is tuned through resonance will not be of equal magnitude. It will be shown later that the frequency read on the wavemeter scale when the meter is so adjusted as to split the silent interval in the beat note will not be the true reading if the maxima mentioned above are of unequal magnitude. If the frequencies under measurement are very high this error will be negligible, but at low frequencies it may be appreciable. It will also be shown that errors of this type are reduced if the wavemeter is coupled to the oscillator, the frequency of which is desired, rather than to the auxiliary oscillator.

If the coupling between the wavemeter and the oscillator is too close no silent mid-point in the beat note can be obtained but as the wavemeter is tuned through resonance the beat note will first increase to a maximum as before, and will then fall off slightly but instead of passing sharply through zero it will jump suddenly to another frequency and then fall gradually to zero. When the jump occurs resonance has been passed. This state of affairs should be avoided. The significance of this frequency jump is discussed in Part 2.

The resonant setting of the wavemeter can be determined with a degree of sharpness that is very great. It is superior in this respect to the *grid dip* method of resonance indication, in which a sensitive galvanometer is included in the grid circuit of the oscillator and is of course vastly superior to the methods which involve the actuation of an indicating device in the wavemeter circuit.

We shall now discuss the theory of the method from the point of view of the equations of a typical vacuum-tube oscillator.

Part 2—Theory of the Method

We wish to study the variations of the beat frequency as the natural frequency of the wavemeter is altered. As a starting point we shall consider the conditions of oscillations of a typical vacuum-tube circuit, such as is shown in Fig. 1. The condition which determines the frequency of oscillation is that the effective reactance of the grid circuit shall be zero. This condition[2] is as follows:

$$X_2+hX_p-\frac{M^2\omega^2X_1}{R_1^2+X_1^2}=0 \tag{1}$$

In which $X_1=\omega L_1-1/\omega C_1$ $\qquad X_2=\omega L_2-1/\omega C_2 \quad X_p=L\omega$

$$h=\frac{\mu m/C_2-m^2\omega^2}{Z_p^2} \qquad Z_p^2=(R+R_p)^2+L^2\omega^2$$

If m and C_2 are kept constant, then h, the coefficient of regeneration, is practically constant, and will be so considered here. This

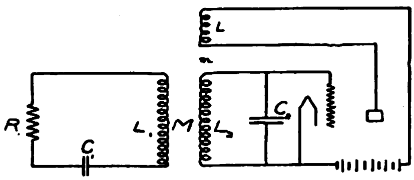

Fig. 1

amounts to assuming that $\mu/\omega C_2 \gg \omega m$ and that $\omega^2L^2/(R+R_p)^2 \ll 1$, which is usually the case in practice. Now we have

$$X_1=(\omega^2L_1C_1-1)/\omega C_1=L_1(\omega^2-\omega_1^2)/\omega$$

$$\omega_1=1/\sqrt{L_1C_1}$$

$$X_2+hX_p=\omega(L_2+hL_p)-1/\omega C_2$$

$$=(\omega L_2'-1/\omega C_2)=L_2'(\omega^2-\Omega_2^2)/\omega$$

$$\Omega_2=1/\sqrt{L_2'C_2}$$

Inserting these relations in (1), gives

$$\frac{L_2'}{\omega}(\omega^2-\Omega_2^2)-\frac{M^2\omega^2}{R_1^2+\dfrac{L_1^2}{\omega^2}(\omega^2-\omega_1^2)^2}\cdot\frac{L_1}{\omega}(\omega^2-\omega_1^2)=0 \tag{2}$$

[2] See p. 339 of "Regeneration in Coupled Circuits," by E. L. Chaffee. Proc. I. R. E., June 1924.

Since Ω_2 is a constant it is possible to obtain ω as a function of ω_1 and this relation would yield the desired information. However, the above equation is of the third degree in ω^2 and of the second degree in ω_1^2 and a direct treatment leads to extreme complications.

More convenient results can be obtained by solving for $(\omega-\Omega_2)$ as a function of $(\omega_1-\Omega_2)$ and introducing certain well-justified approximations. Equation (2) can be written

$$\frac{L_2'}{\omega}(\omega-\Omega_2)(\omega+\Omega_2)-\frac{M^2\omega^2\frac{L_1}{\omega}(\omega-\omega_1)(\omega+\omega_1)}{R_1^2+\frac{L_1^2}{\omega^2}(\omega-\omega_1)^2(\omega+\omega_1)^2}=0$$

Let $y=\omega-\Omega_2$ and $u=\omega-\omega_1$

$$yL_2'\left(1+\frac{\Omega_2}{\omega}\right)-\frac{M^2\omega^2L_1u\left(1+\frac{\omega_1}{\omega}\right)}{R_1^2+L_1^2u^2\left(1+\frac{\omega_1}{\omega}\right)^2}=0 \qquad (3)$$

Now assume that $\Omega_2/\omega=\omega_1/\omega=1$ and call $\Omega_2=\Omega=\omega$

Then $$2L_2'y-\frac{2M^2\Omega^2L_1u}{R_1^2+4L_1^2u^2}=0 \qquad (4)$$

Let $$x=\omega_1-\Omega=y-u$$

Eliminating u from (4) gives

$$L_2'y-\frac{M^2\Omega^2L_1(y-x)}{R_1^2+4L_1^2(y-x)^2}=0 \qquad (5)$$

or $4L_1^2L_2^1(y^3-2y^2x+yx^2)+(R_1^2L_1^2-M^2\Omega^2L_1)y+M^2\Omega^2L_1x=0.$
This can be written $y^3-2y^2x+yx^2+Dy+Ex=0$ (6)
In which

$$\left.\begin{aligned} D&=\frac{R_1^2L_2'-M^2\Omega^2L_1}{4L_1^2L_2'}=\frac{(\eta_1^2-\tau^2)\Omega^2}{4}\\ E&=\frac{M^2\Omega^2L_1}{4L_1^2L_2'}=\frac{\tau^2\Omega^2}{4}\end{aligned}\right\} \qquad (7)$$

$$\left.\begin{aligned} D+E=\frac{\eta_1{}^2\Omega^2}{4}>0 \\ \eta_1=\frac{R_1}{\Omega L_1} \qquad \tau=\frac{M}{\sqrt{L_1L_2'}} \end{aligned}\right\} \tag{8}$$

$y/2\pi=(\omega-\Omega)/2\pi$ is the beat frequency and $x/2\pi=(\omega_1-\Omega)/2\pi$ is the frequency difference between the wavemeter setting and the resonant frequency. E cannot be less than zero but D may be positive, negative, or zero.

The obtaining of y as a function of x necessitates the solution of a cubic equation and involves calculations of considerable complication. We shall examine the equation and obtain all the information we desire without actually solving it.

Let us note that the substitution of $-x$ for x or of $-y$ for y in (6) reduces it to

$$y^3+2y^2x+yx^2+Dy-Ex=0 \tag{9}$$

Since (6) and (9) are not identical it follows that the curve $y=f(x)$ is not symmetrical with respect to either the x or y axes. However, the simultaneous substitution of $-x$ for x and $-y$ for y leaves (6) unchanged and hence $y=f(x)$ is symmetrical with respect to the origin. Equation (6) may be written

$$\frac{Ex}{y}=-\left[(y-x)^2+D\right] \tag{10}$$

Now $(y-x)^2\geqq 0$. Hence if D is positive x/y is always negative and $y=f(x)$ lies entirely in the second and fourth quadrants and must either be discontinuous or pass through the origin. A detailed investigation shows that the curve is continuous at all points. If D is negative the curve lies in the first and third quadrant for small values of $(y-x)^2$ and in the second and fourth for values of $(y-x)^2$ which are larger than $-D$. When $x=0$ then $y^3+Dy=0$ by (6). Hence

$$y=0 \text{ or } y=\pm\sqrt{-D} \tag{11}$$

If $D\geqq 0$ then $y=0$ is the only solution, while if $D<0$ there are three possible values of y corresponding to $x=0$, that is, to the resonant setting of the wavemeter.

Solving (6) for $x=F(y)$ we obtain

$$x=y-E/2y\pm\sqrt{E^2/4y^2-(E+D)} \tag{12}$$

If x is to be real there is imposed upon y the restriction that

$$y^2 \leqq E^2/4(E+D) \tag{13}$$

$$\frac{dx}{dy}=1+\frac{E}{2y^2}\left(1\mp\sqrt{1-\frac{4y^2(E+D)}{E^2}}\right)\mp\frac{\frac{2(E+D)}{E}}{\sqrt{1-\frac{4y^2(E+D)}{E^2}}} \tag{14}$$

If we choose the positive sign in the ambiguity then when $y=0$, $dx/dy=\infty$. This corresponds to $x=\pm\infty$. If we choose the negative sign, then when $y=0$, (14) is an indeterminate form, the evaluation of which gives $dx/dy=-D/E$. Hence the slope, at the origin, of $y=f(x)$ is negative when $D>0$, is infinite when $D=0$ and is positive when $D<0$. dx/dy is also infinite when $1-4y^2(E+D)/E^2=0$. This corresponds to a maximum value of y.

$$y_m=\pm\frac{E}{2\sqrt{E+D}} \tag{15}$$

Inserting (15) in (12) we obtain the value of x which corresponds to the maximum value of y

$$x_m=\mp\frac{2D+E}{2\sqrt{E+D}} \tag{16}$$

We are now in possession of the following information:

(a) $y=f(x)$ is not symmetrical with respect to either the x or the y axes but is symmetrical with respect to the origin.

(b) For $D>0$, f lies entirely in the second and fourth quadrants. For $D<0$, f lies in the first and third quadrants for $(y-x)^2<|D|$ and in the second and fourth for $(y-x)^2<|D|$.

(c) The curve passes through the origin.

(d) y is restricted in magnitude by the relation $y^2\leqq\dfrac{E^2}{4(E+D)}$

(e) Within the range of (d) there are two distinct values of x for every value of y except for $y=\pm E/2\sqrt{E+D}$ when the two values of x are identical.

(f) The maximum and minimum values of y are given by the relation of (d). The values of x corresponding to these values of y are given by $x_m=\mp\left(\dfrac{2D+E}{2\sqrt{E+D}}\right)$

(g) The slope of the curve at the origin is given by

$$\frac{dy}{dx}=\frac{-E}{D}=-\frac{\tau^2}{(\eta_1^2-\tau^2)}$$

In Fig. 2 are plotted three curves for $\Omega=4\times10^6$ or $\lambda=472$ meters, $\eta_1=0.004$ and $\tau=0.002$, $\tau=0.004$ and $\tau=0.006$ for curves 1, 2, and 3 respectively.

As the resonant frequency of the wavemeter is increased from a value smaller than $\frac{\Omega}{2\pi}$, the beat frequency y, is seen to increase slowly, reach a maximum, and then to drop off sharply as resonance is approached. This falling off of y will be the more rapid, the larg-

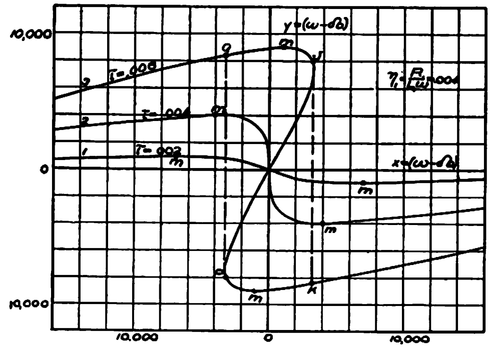

Fig. 2

er the value of τ, up to $\tau=\eta_1$. If $\tau>\eta_1$ then y will not pass through zero but after reaching its maximum value will decrease somewhat to a point such as J of curve 3. As ω_1 is further increased a jump to the point K takes place and a sudden change in the beat note results. If the order of tuning is reversed and ω_1 is decreased from a value larger than that corresponding to the point K then a jump again occurs but this time it is from P to Q and takes place at a value of ω_1 which is different from that which corresponded to the first jump. It is evident that this state of affairs is undesirable in making frequency measurements, although of course the interval in ω_1 may be split in order to obtain the resonant frequency. The ideal condition for precision measurements is the critical one, $\tau=\eta_1$. The middle region of inaudible beats is then as narrow as it is possible to have it and y_m will be larger than for any other curve

which does not give a discontinuity in the beat frequency. Calling the value of y_m corresponding to $\tau=\eta_1$, y_{mm} we have:

$$y_{mm}=\frac{\sqrt{E}}{2}=\frac{\eta_1\Omega}{4} \tag{17}$$

This is directly proportional to the frequency. On the other hand $\frac{y_{mm}}{\Omega}$ is dependent only on the quantity η_1 which is, for many coils, almost independent of frequency over a considerable range. In the case cited $\frac{y_{mm}}{\Omega}=0.001$ or 0.1 per cent and the approximations $\frac{\Omega}{\omega}=\frac{\omega_1}{\omega}=1$ which were introduced into the original equations, are well justified. It is evident that the method is better adapted to the higher frequencies since the value of $\frac{y_m}{2\pi}$ must be in the audible range, and if the frequency under measurement is low, y_m may be too large a fraction of Ω. However, the range of applicability may be extended several octaves by adjusting the auxiliary oscillator to zero beats with a harmonic of the oscillator which is being measured. If one of the higher harmonics is thus employed it may be necessary to insert the telephones in the circuit of the oscillator under measurement in order that the beat note may be audible without excessive amplification. Frequencies outside of the range of the wavemeter may also be determined by employing the harmonics of one or both oscillators.

If the two oscillators are not exactly in resonance a certain error will be introduced into the determination of the wavelength. Suppose that the frequency $\frac{\Omega_3}{2\pi}$ of the auxiliary oscillator is slightly lower than $\frac{\Omega}{2\pi}$, the frequency which is being determined. If the interval ab, in Fig. 3, represents the magnitude of the minimum frequency of audibility, then when the wavemeter is absent the beat

note $\dfrac{\Omega_3-\Omega}{2\pi}$ will be inaudible. When the wavemeter is coupled to the oscillator which, when uncoupled to other circuits, has a frequency of $\dfrac{\Omega}{2\pi}$ and the meter is tuned through resonance, the beats will become audible at the points C and C' and the reading which is obtained will be too high by the amount $\Delta\omega = OS$, where S is the point which is judged to be the center of the interval CC'. The difference between the wavemeter reading and the frequency of

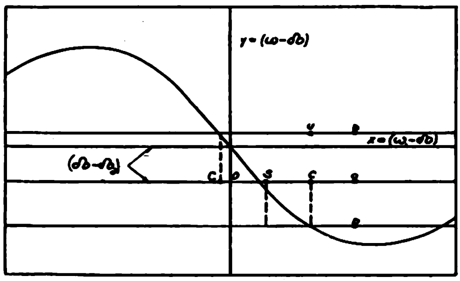

Fig. 3

the auxiliary oscillator is even greater, being $(\Omega_3-\Omega)+OS$, and hence it is apparent that the wavemeter should be coupled to the oscillator, the frequency of which is to be measured. There is also evident the importance of having the two oscillators accurately adjusted to zero beat. When the beat note is inaudible but is not zero it will be noted that the maximum beat frequency will be greater when the wavemeter is tuned away from resonance in one direction, then in the other. By adjusting the auxiliary oscillator until these maxima are judged to be equal the errors from this source may be practically eliminated.

The frequency of a non-oscillating circuit can best be found by obtaining sufficient data for a calibration curve. The two oscillators are adjusted to zero beat at a frequency somewhere near that which the non-oscillating circuit is thought to have. The latter is then loosely coupled to one oscillator and its capacity varied until the silent mid-point is found. The circuit is then removed and a wavemeter put in its place. Without changing the adjustments of the oscillators the wavemeter is brought into resonance. The frequency of the wavemeter circuit is then the

same as that of the circuit under measurement. By obtaining several such points a calibration curve can be drawn and the frequency corresponding to any given condenser setting may be obtained by interpolation. If the circuit elements are invariable an auxiliary condenser may be added and two or three points obtained. If these points are properly chosen a fairly accurate extrapolation may be made to find the frequency which is associated with the circuit when the auxiliary condenser is removed.

$$\text{The relation } \tau = 2\sqrt{\frac{y_m \eta_1}{\Omega}} \text{ or } M = \frac{2}{\Omega}\sqrt{y_m R_1 L_2} \tag{15a}$$

suggests a possible method for the measurement of the coefficient of coupling or the mutual inductance. If the mutual inductance between two circuits is to be measured, one of them, the inductance of which is known, may be connected to a vacuum tube and made to function as one of the two oscillating circuits required to obtain zero beats while the other replaces the wavemeter circuit in the foregoing discussions. If the various circuit constants are such that $\frac{y_m}{2\pi}$ lies within the audible range it may be measured on a frequency meter of the resonance bridge or other suitable type. Then if R_1, L_2 and Ω are known, M may be determined from (15a). The variation of y with x in the neighborhood of y_m is small and hence there should be no difficulty involved in setting x so as to obtain a value of y which is very near the maximum. R_1 can be found by any of the standard methods for the determination of resistance at high frequencies. Ω must be determined by means of a wavemeter as in the above. Since y_m, R_1 and L_2 enter under the radical sign the per cent error introduced into the value of M by a small error in any one of these quantities will be half the per cent error in that quantity.

In case a frequency meter is not available y_m may be adjusted to zero beat with a tuning fork by inserting a variable resistance in the circuit (1) and changing R_1 until y_m has the proper value.

Another method which might suggest itself as applicable in case there is no frequency meter to be had is that of bringing y into zero beat with a tuning fork by varying x. This adjustment is possible for four values of x provided that the frequency of the fork is less than y_m. Since y now differs from y_m the relation

between y and τ will contain x and (12) must be used. By substituting two pairs of corresponding values of x and y in (12) we obtain two simultaneous equations in x and y. If we choose two pairs that are symmetrical with respect to the origin the resulting equations are identical and cannot be solved. If we choose two pairs such that both values of x lie on the same side of the origin the sign of the ambiguity in (12) must be different in the two equations. On subtracting we obtain

$$x_2-x_1=2\sqrt{\frac{\tau^4\Omega^4}{64y^2}-\frac{\eta_1{}^2\Omega^2}{4}} \tag{18}$$

or

$$\tau=\frac{2}{\Omega}\sqrt[4]{y^2(x_2-x_1)^2+y^2\eta_1{}^2\Omega^2} \tag{19}$$

In which x_1 and x_2 lie on opposite sides of y_m and are values of x which give identical values of y.

If we choose y sufficiently small (x_2-x_1) will be large and we may fulfil the condition

$$(x_2-x_1)^2>>\eta_1{}^2\Omega^2=\left(\frac{R_1}{L_1}\right)^2 \tag{20}$$

In this case

$$\tau=\frac{2}{\Omega}\sqrt{y(x_2-x_1)} \tag{21}$$

Because of the extreme flatness of the curve for large values of x there will be a large probable error in determining x_2, the larger value of x. In order that the condition (21) may be fulfilled it is necessary that x_2 be extremely large. Because of the asymptotic approach of y to zero as x increases the difficulties involved in determining the value of x_2, which corresponds to a given value of y, are considerable and consequently this method cannot be considered as a satisfactory one.

In conclusion I wish to acknowledge my indebtedness to Professor G. W. Pierce of Harvard University, under whose direction I first used this method of frequency measurement, for permission to work up and discuss the theory involved.

PRECISION DETERMINATION OF FREQUENCY*

By

J. W. Horton and W. A. Marrison

(Bell Telephone Laboratories, New York City)

Summary—The relations between frequency and time are such that it is desirable to refer them to a common standard. Reference standards, both of time and of frequency, are characterized by the requirement that their rates shall be so constant that the total number of variations executed in a time of known duration may be taken as a measure of the rate over shorter intervals of time. Frequency standards have the further requirement that the form of their variations and the order of magnitude of their rates shall be suitable for comparison with the waves used in electrical communication.

Two different types of standard which meet these requirements are described. One consists of a regenerative vacuum-tube circuit, the frequency of which is determined by the mechanical properties of a tuning fork. The other is a regenerative circuit controlled by a piezo-active crystal. Means are provided, in the case of each standard, whereby the recurrent cycles may be counted by a mechanism having the form of a clock, the rate of which is a measure of the frequency of the reference standard.

Data taken over a period of several years with a fork-controlled circuit show that, under normal conditions, its rate may be relied upon to two parts in one million. Data taken over a much shorter time with crystal controlled oscillators indicate that they are about ten times as stable.

WHEN we speak of the frequency of a wave or, in fact, when we use the word frequency in any connection, we mean, in general, the number of times a periodic occurrence takes place within a given interval of time. Consequently the frequency of a recurrent phenomenon is the reciprocal of the time interval required for the variation to pass through one cycle provided the duration of any given cycle is identical with that of any other, or, in other words, that the variation has a constant frequency. Since, therefore, frequency may be expressed completely in terms of time, it is neither necessary nor desirable to have any fundamental standard of frequency other than the accepted standard time interval.

The present standard of time is, of course, the sidereal day, which is the time required for the earth to make one complete revolution in space. A succession of sidereal days, therefore, constitutes a recurrent phenomenon the duration of any cycle of which is, by definition, equal in time to the duration of any other

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cycle. In other words, the variation has the essential characteristics required of a standard of frequency. The mean solar day, in terms of which all measurements of time are expressed, must be considered as a practical standard the determination of which is made in terms of the sidereal day.

There is a unique difference between a standard of time—or frequency—and standards of most other quantities. For example, the standard unit of mass or the standard unit of length may be given physical embodiments and may be put in a safe where their identity will be preserved over extended periods. Unlike a mass or a length, an interval of time cannot of itself be preserved nor can it readily be given a physical embodiment which by virtue of its great constancy may be used as a fundamental standard. At present, therefore, all precision measurements of time must be referred directly to the rate of the earth's rotation.

Having once concluded that the fundamental standard of time is, *ex officio*, the fundamental standard of frequency, there remains simply the problem of comparing the rate of any periodic variation which is to be measured with the rate of the earth's rotation.

If there were available only the periodic phenomenon of unknown rate and the intervals defined by the earth's rotation, it would be necessary that the unknown phenomenon maintain its rate unchanged throughout an entire interval covered by one rotation of the earth. In this case the measurement would consist in counting the number of times the unknown variation repeated itself during the interval. Due, however, to the availability of means for accurately defining time intervals less than the sidereal day to the required precision, it is, of course, unnecessary to maintain the rate of the variation to be measured over the longer interval. When we recognize that any means used for evaluating the time of short intervals is itself executing periodic variations of equal duration, we realize that it is a secondary standard of frequency as well as a secondary standard of time. Thus the Riefler clocks at the National Bureau of Standards and at the U. S. Naval Observatory become our most practical working standards of frequency. In terms of the oscillations of their pendulums their frequency may be said to be unity when referred to the second.

A secondary standard such as a high grade seconds pendulum is unsuited to the actual determination of the frequency of such variations as are of interest in electrical communication, both

because of the excessively high ratio between the rates of the occurrences to be compared and because, in its usual form, the motion of a clock pendulum is not well suited for comparison with electrical variations. It is desirable, therefore, as the first step in the precision measurement of the frequencies of electric waves, to provide a suitable reference standard. The requirements of such a standard are that its rate should be of the same order of magnitude as the rates to be measured, that the form of the variation which it controls should be suitable for convenient comparison with the variations to be measured, and, finally, that its rate shall be sufficiently constant so that the total number of variations which it executes in a time of such duration that it may be defined with high precision may be taken as the rate of the variation over shorter intervals of time. This last requirement again emphasizes the inherent similarity between secondary standards of frequency and secondary standards of time. For convenience, it should also be possible to adjust the rate of the frequency standard to some prescribed value within such limits that the error may be neglected in the majority of measurements.

Original Equipment

A secondary standard having the characteristics mentioned above was described several years ago by the authors of the present paper in collaboration with Mr. N. H. Ricker. This standard consists of a 100-cycle tuning fork maintained in vibration by an amplifier regeneratively connected through the fork by electromagnetic coupling. Means are provided for obtaining from the electrical portion of the resulting oscillating system a sinusoidal alternating current the frequency of which is constant to high precision. Standards of this general type are being extensively used, as will be seen by referring to the bibliography. A brief survey of the present status of this 100-cycle standard is of interest not only because of the improvement in its performance since the original report was presented, but also because the experience which has been gained from its use has been of value in the development of standards of still greater utility.

The fork which was described in the paper already referred to ran continuously from April, 1923, to May, 1927, except for four intervals totalling about three days. Of these interruptions two were from accidental causes and two for the purpose of making minor changes in the system. Throughout this entire period the

temperature of the fork, as maintained by its thermostatically controlled water bath, was held at approximately the value for which its rate is 100 cycles. Since the temperature coefficient of frequency for the particular fork in question is 0.0109 per cent per degree Centigrade, it is necessary to keep the temperature to within 0.01 deg. C. of the prescribed value in order that the rate shall be correct to one part in a million. It has been found that the temperature coefficient, instead of being a detriment,

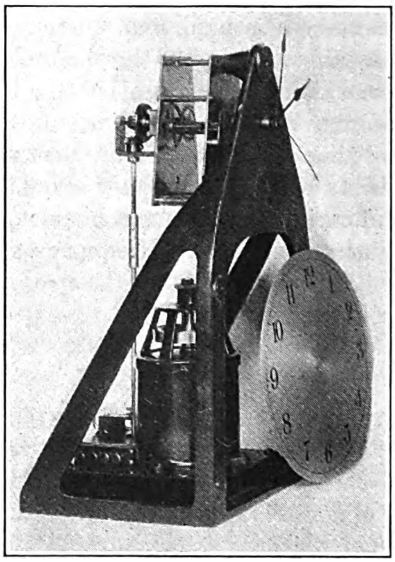

Fig. 1—100-cycle Synchronous Motor Geared to Clock Train.

has furnished an excellent means for adjusting the frequency, making possible small changes the amounts of which may be accurately predetermined. During the four years of operation, the change in the frequency of the entire system due to aging of the fork and to all other causes, as determined by the change in the temperature—indicated by a Beekman thermometer—at which the rate is exactly 100 cycles, has been less than 0.004 per cent.

The synchronous motor described in the previous paper has been replaced by an improved form shown in Fig. 1. The original motor was coupled to the clock by means of a gear reduction driving a commutator, from which seconds impulses were supplied to a clock train having an electromagnetic stepping device. The present motor is geared directly to the clock train, thus completely

avoiding errors due to the mechanism. In spite of the lack of confidence expressed by many people in the impulse type of synchronous motor, both of those mentioned have been found to give entirely satisfactory operation over long periods of time. In fact, the only occasions on which the motors have ever stopped in service have been those on which power was taken off some portion of the system.

During tests on this frequency standard, it was found that it constituted a far more reliable timekeeper than the electrically maintained pendulum clock which was used to obtain the data already published. The pendulum clock was, therefore, dispensed

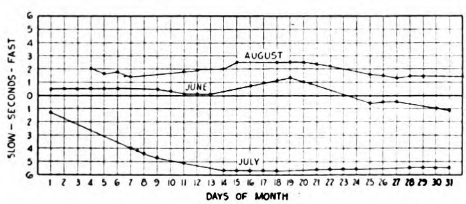

Fig. 2—Daily Error in Clock Controlled by 100-Cycle Frequency Standard. Mean rate for three month period 99.99991 cycles; average deviation from mean rate 0.00016 cycles; maximum deviation from mean rate 0.00067 cycles.

with and all measurements of the rate of the fork are now made by direct comparison with the mean solar day as defined by the radio time signals sent out from the U. S. Naval Observatory. By means of its second hand, which is concentric with the hour and minute hands and which rotates uniformly, the daily error in the fork-controlled clock may be determined to 0.2 second, giving its average rate for the day to approximately three parts in one million. For greater accuracy there is provided, in the gear train between the motor and the clock, a shaft rotating once per second and operating a contact by means of a cam. Pulses from this contact may be recorded simultaneously with the radio time pulses on a moving paper tape by means of a two-element siphon recorder. From the record thus obtained, the error of the clock may be determined to 0.01 second, increasing the accuracy of the determination about 20 times.

The data shown by the chart of Fig. 2 are typical of the performance of this system. On this chart the error in the clock reading is plotted against days. The rate is obtained from the slope of the curve; when it is straight the rate is constant; when

it is horizontal the rate is exactly 100 cycles. Attention is called not only to the constancy of the rate, which at no time during the period covered by this record departed more than 0.0007 per cent from the mean rate, but also to the precision with which the system was adjusted to the desired absolute value.

A method for the comparison of high frequencies with the 100-cycle standard is also outlined in the previous paper. This method involves the use of harmonic producers and of harmonic selecting networks. By repeated harmonic production and selection it is possible to obtain any multiple of the fundamental frequency to a limit determined by the care taken. Theoretically the process may be extended indefinitely, but practically there are difficulties in going beyond two or three stages. These difficulties arise from small irregularities in the frequency or in the amplitude of the current in the initial stages of the system, all of which result in frequency irregularities in the final wave. While these may be entirely negligible in the wave in which they originate, they are enormously magnified in subsequent stages. With ordinary care it is feasible to obtain from the 100-cycle standard current a current having a frequency of 100,000 cycles per second or even higher. By the use of a cathode ray oscillograph, an independent oscillator may be adjusted so that its frequency is an exact multiple of the frequency of the current thus obtained. This process may be repeated by successive stages up to the limit of the oscillograph tube.

The system, therefore, contains three essential elements: (1) The regeneratively driven fork constitutes a source of alternating current, the frequency of which is 100 cycles to a very high precision. (2) The synchronous motor and the gear train of the clock constitute a rate reducing mechanism having a ratio of approximately ten million to one. (3) The harmonic producer system and the cathode ray oscillographs may be operated so as to increase the frequency in the ratio of several thousand to one. It is, therefore, practicable by means of this system to determine the frequency of an electric wave, using the mean solar day as a standard, when the ratio between them does not exceed about 10^{11}.

Present Equipment

At the time the development of the 100-cycle standard was undertaken, the frequency range over which measurements were required to high precision did not exceed several hundred thousand cycles per second. For this work the frequency chosen for the

secondary standard was quite satisfactory. Since then, however, more and more attention has been given to the higher frequencies until it has become necessary to make measurements to a very high degree of accuracy up to several million cycles per second. Any appreciable increase in the frequency of the reference standard is obtained at the expense of an increase in the ratio between its rate and the mean solar day. This is not particularly objectionable, however, since the comparison between these two rates may be made with permanently adjusted apparatus designed expressly for this fixed ratio, whereas comparisons between unknown rates and the reference standard must be made by carefully adjusting

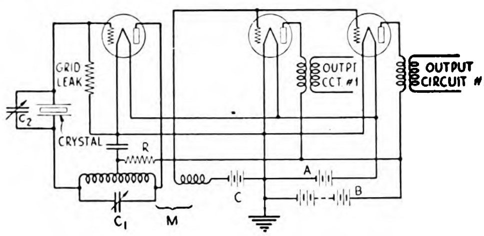

Fig. 3—Circuit of Oscillator for High Frequency Standard

apparatus capable of working over a wide range of ratios. More important still, it has been found that the frequency of a given current can be controlled by a current of higher frequency with much greater stability than by a current of lower frequency. Because of these various factors, therefore, there is need for a standard of the general type just described but having a fundamental frequency many times higher. Such a standard has been developed and some preliminary data as to its constancy of rate have been obtained.

The high-frequency generator of this standard is a 50,000-cycle quartz crystal-controlled oscillator having circuit elements and controls that give a high degree of stability. The circuit of this oscillator is shown in Fig. 3. It is of the Hartley type, differing from the usual form only in having a resonant piezo-electric crystal connected in the grid circuit. This type of circuit is used because it is convenient, by means of circuit adjustments, so to control the phase of the regenerative feedback that maximum stability is obtained. Being an efficient oscillating circuit it also permits of using the crystal loosely coupled to the electrical circuit, thus permitting full advantage to be taken of the very low decrement of the crystal.

The crystal is adjusted by lapping so that the frequency of the circuit controlled by it is 50,000 cycles exactly at a given operating temperature. Minute corrections that are subsequently required can be made by means of a small variable condenser in parallel with the crystal.

The crystal-controlled oscillator is coupled to two output circuits through vacuum tubes, having their grids in parallel, loosely coupled to the tuning coil. In this way considerable output

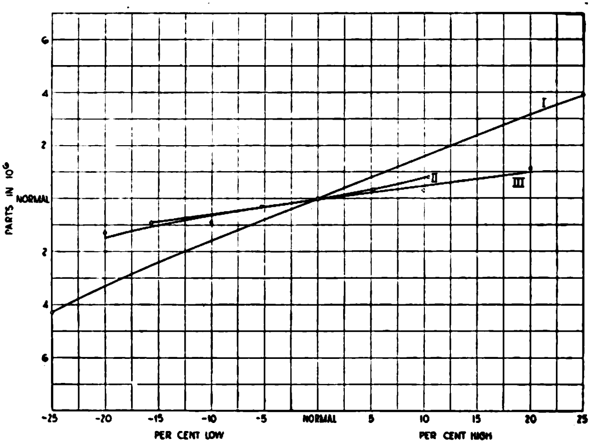

Fig. 4—Effect of Circuit Variations on Frequency of 50,000-cycle Standard Oscillator. I—Change in frequency vs. percentage change in the grid leak resistance; II—Change in frequency vs. percentage change in filament current; III—Change in frequency vs. percentage change in plate voltage.

current at 50,000 cycles can be obtained with no possibility of reaction on the oscillator from external circuits.

The three curves in Fig. 4 show the frequency stability of the oscillator under variations in filament current, plate voltage, and grid leak resistance. The departure from normal frequency in parts in a million is plotted against departure in per cent from normal filament current, plate voltage, and resistance.

The variation with grid leak resistance is greatest, but due to the use of a special type of resistance that is expected not to vary as much as 0.1 per cent in service, the variations in frequency due to this source are negligible.

The variation with battery voltage is the most serious factor, being in the same sense for both batteries. These voltages are controlled at present to about ± 2 per cent, therefore variations in frequency of about three parts in ten million may be expected

due to them. Of course, as has been done in other cases, it is possible to compensate the circuit for variations in voltage so that any tendency for a change in frequency in one direction is offset by an equal and opposite tendency.

The curves of Fig. 5 show the variations in frequency of the oscillator for changes in the condensers C_1 and C_2. These curves show variations in frequency in parts in a million and are plotted, in the case of condenser C_2, against dial settings, and in the case of condenser C_1, against the capacity in micro-microfarads. Due

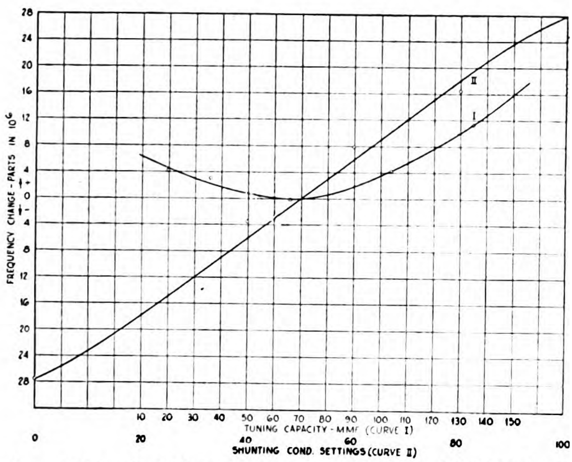

Fig. 5—Effect of Circuit Adjustments on Frequency of 50,000-Cycle Standard Oscillator. I—Change in frequency vs. capacity of tuned circuit of oscillator; II—Change in frequency vs. second of condenser shunting crystal.

to the linearity of the adjustment obtained by condenser C_2 it is possible to make predetermined minute changes in the operating frequency. The proper adjustment for condenser C_2 is very nearly that value for which the change in frequency with capacity is zero. Under this condition the small changes in the electrical circuit due to temperature variations have an entirely negligible effect on the frequency.

The quartz crystal in its mounting is shown in Fig. 6. It vibrates in the direction of its length and is supported on a short piece of felt at the center in order to allow free vibration. It is held in place and separated from the plates by silk threads. The logarithmic decrement of this crystal in its mounting, the whole being measured as an electric circuit, is about 0.00012. The conductor plates are made very rigid in order to keep the separation constant, since a variation in this spacing as small as 0.01 mm.

would cause an appreciable error in frequency. The temperature coefficient of frequency is about 0.00038 per cent per degree Centigrade, requiring a control of temperature constant to within 0.025 deg. C. in order to keep variations due to temperature change alone within one part in ten million. Measurements on the effect

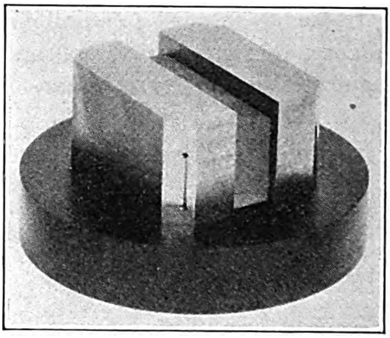

Fig. 6—Resonant Piezoelectric Quartz Plate in Mounting.

of atmospheric pressure, covering the range from 760 mm. to very low pressures, show the frequency to be approximately a linear function of the pressure. For the particular crystal measured, it was found that an increase in pressure of 140 mm. caused a decrease in frequency of 0.001 per cent.

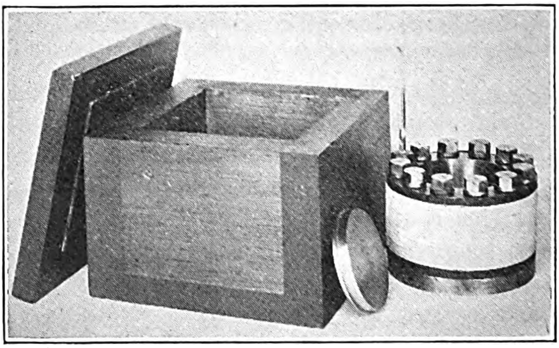

Fig. 7—Thermostat for Piezoelectric Plate.

The thermostat used at present to control the temperature of the crystal is shown in Fig. 7. It consists of a steel cylinder with hollow walls filled with mercury which expands when heated into a capillary tube, as in a thermometer. At a certain tempera-

ture contact is made on a pointed tungsten wire (not shown) which operates a mechanism for changing the rate of heating applied through a resistance coil wound on the outside of the steel cylinder. The crystal is mounted within the cylinder and the whole device is enclosed in a balsa-wood box for heat insulation. By means of this arrangement the temperature at a point within the cylinder may be kept at a prescribed value over long periods.

It is of interest to note that the temperature of the crystal when operating is not exactly that of the space surrounding it because some energy is dissipated within the crystal. For this

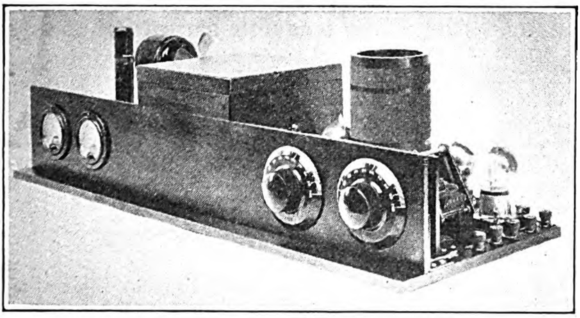

Fig. 8—Standard 50,000-cycle Oscillator.

reason, since the dissipation varies at least as the square of the applied plate voltage, that voltage is kept at a low value. The resistance R in the circuit of Fig. 3 is for this purpose.

The vacuum-tube circuit of the oscillator with the crystal and temperature control apparatus is shown in the photograph Fig. 8. This whole circuit is mounted in a closed cabinet which itself is temperature controlled by a commercial thermo-regulator.

An important element of this new frequency standard is the submultiple frequency generator which is used as the first element of the system for comparing the rate of the crystal-controlled oscillator with the time standard. This submultiple producer consists of an oscillator for generating the low frequency, a harmonic producer for obtaining a harmonic of the low frequency which corresponds to the high frequency by which it is controlled, and a modulator from which is obtained a direct current the amplitude of which is a function of the phase relation between the controlling high-frequency current and the harmonic of the controlled low-frequency current. The action of the direct current

on a control element in the low-frequency oscillator maintains the frequency at such a value that the current from the harmonic producer has exactly the same frequency as that from the standard oscillator. Fig. 9 shows the circuit arrangement. The tubes, from the left, are associated with the modulator, the oscillator, the harmonic producer and an output amplifier, respectively.

The control device consists of a core of magnetic material having two windings, one in the oscillating circuit proper and one in the plate circuit of the modulator tube. This coil is designed so that a variation in the direct current output of the modulator causes a variation in the inductance of the winding in the oscillating circuit by virtue of the magnetic saturation of the core.

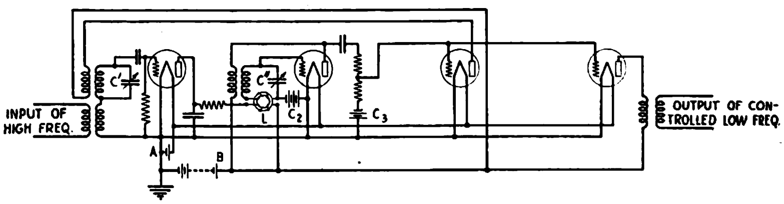

Fig. 9—Submultiple Controlled Frequency Generator

In operating, the low-frequency oscillator is adjusted so that the frequency is exactly some submultiple of the control frequency when the direct current into the control coil has a certain mean value. If, then, anything occurs that tends to change the low frequency, the resulting phase shift between the harmonic and the control current instantly causes a change in the direct current in the control coil that opposes that tendency. The result is that, in spite of large variations in the low-frequency circuit, the frequency is maintained at an exact submultiple of the high-frequency control, the only variation being a slight shift in phase with respect to the control current.

With a circuit operating in this fashion it is possible to control a current so that its frequency shall be maintained in an exact integral relationship to that of some other current, even when the latter is many times higher. A ratio of 50 to 1 may be readily secured with moderate care. It would thus be entirely feasible to secure, by repeated stages, a current having a frequency which was exactly one five-hundredth the frequency of the 50,000-cycle standard, or 100 cycles. This current could then be used to run the synchronous motor of the clock previously described. Actually

it has been found advantageous to secure a current having a frequency one-tenth that of the standard and to use this current to operate a 5000-cycle synchronous motor.

This high-frequency synchronous motor operates on the output of a single 5-watt tube. The rotor is mounted on a vertical steel shaft running on a single steel ball under oil. The motor is mounted under a bell jar and is operated in partial vacuum in order to reduce friction losses and to keep the bearings clean. It takes more than half an hour for the motor to come to rest from full speed when it is disconnected from the source of power.

The rotor is a flat disk having 100 teeth milled on one side for poles and having an annular recess in the upper side for the

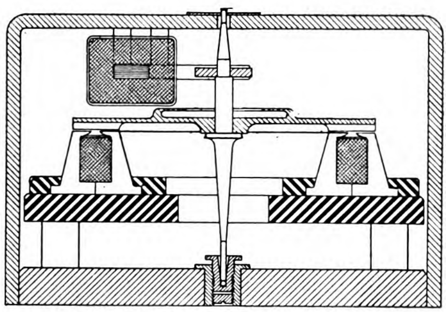

Fig. 10—Sectional View of 5,000-cycle Synchronous Motor.

mercury used to prevent hunting. The stator consists of 100 U-shaped pieces of silicon steel arranged in radial planes equally spaced around the axis. These form a ring of the same outside diameter as the rotor, and having an annular space in which is located a coil for magnetizing all the poles in parallel. A single U-shaped piece and a single fin on the rotor constitute one magnetic circuit around the coil. By this construction the minimum magnetic material is used and it is possible to laminate properly every part. Coils can be wound to match the electrical impedance of any suitable vacuum tube and can be exchanged easily. The spacing between the rotor and stator can be altered for any special case by adjusting the lower bearing. Figs. 10 and 11 show a sectional drawing and a photograph of the assembled motor.

The coil of the motor is connected in the plate circuit of the tube used to drive it so that both alternating and direct current flow in the winding. This produces 5000 magnetic pulses per second instead of 10,000 as would be the case if alternating current

alone were used. Thus the rotor moves one pole per complete cycle and revolves at 50 revolutions per second.

Maximum power is obtained in the motor if its impedance is conjugate to the impedance of the output circuit of the amplifier. The internal output impedance of a tube is nearly pure resistance, so the reactive component of the motor impedance is balanced by a series condenser between the motor winding and the filament. The space current is supplied through a choke coil connected to the junction of the motor winding and the condenser. A somewhat better arrangement would be to resonate the motor winding by

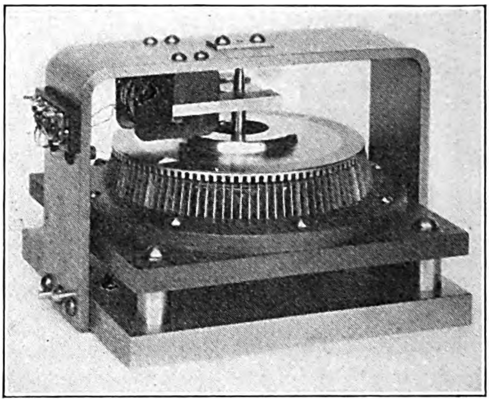

Fig. 11—5,000-cycle Synchronous Motor.

a parallel connected condenser and to connect the plate battery at a tap in the winding. In this way the choke coil can be dispensed with and the ratio of alternating to direct current in the winding can be adjusted so that the resultant goes to zero once during each cycle. At the same time the impedance of the tube can be matched exactly, as required for maximum power transfer.

It would be a simple matter to arrange a gear train so that this motor might operate the clock mechanism directly. Since, however, there were already available clocks designed to operate on a 100-cycle current, it was thought to be simpler to use these as they were. Consequently, the high-frequency motor drives a small generator, the two-pole rotor of which is mounted on the same shaft. Since the speed of the motor is 50 r.p.s., and since the generator is of the inductor type, the generated frequency is 100 cycles—or exactly 1/500th of the original high frequency.

The output of this generator is amplified by a single vacuum tube which supplies sufficient power to operate the motor of the 100-cycle clock, the rate of which, therefore, is an accurate measure of the frequency of the 50,000-cycle oscillator.

In the design of the electrical circuits, provision is made for supplying power from multiple amplifiers at 50,000, 5000 and 100 cycles, which frequencies are in exact harmonic relation. These are available continuously for making frequency comparisons in the three ranges.

In addition to this provision, circuits have been built that permit the control of any frequency which is a multiple or sub-

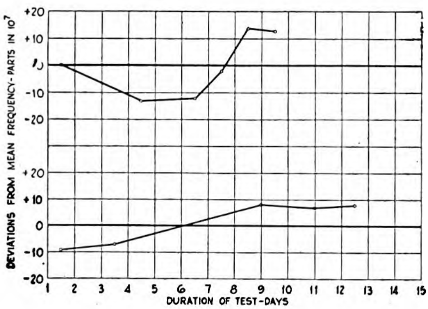

Fig. 12—Daily Variation in Average Frequency of 50-kilocycle Standard.

multiple up to the tenth order of any one of the three original frequencies. Also, many fractional multiples may be thus controlled. This considerably extends the usefulness of the new standard since it is possible to deliver alternating current at constant frequency over a range from 10 cycles to 500,000 cycles per second. These frequencies are not only constant but are known to the same accuracy as the 50,000-cycle control frequency, being some predetermined exact rational multiple or fraction of that frequency.

Because of the short time during which this standard has been in operation, and because of numerous interruptions for the purpose of making slight modifications, data covering a long continuous run are not available. Measurements made over several brief periods are, however, of interest as they indicate the performance which may be expected from the system. Fig. 12 gives the results of two brief runs. On this chart each point represents the deviation of the average rate for a single twenty-four hour interval from the mean rate for the entire period. It is

probable that variations in atmospheric pressure were responsible to a considerable extent for the variations in frequency. It is obviously desirable to provide means whereby the crystal may be operated under constant pressure as well as constant temperature.

The above data indicate a high degree of constancy of the average frequency of this standard over long time intervals. In view of the very great ratio between the rate of the standard and the mean solar day, it is desirable to obtain additional checks as to the constancy of frequency over shorter time intervals. For

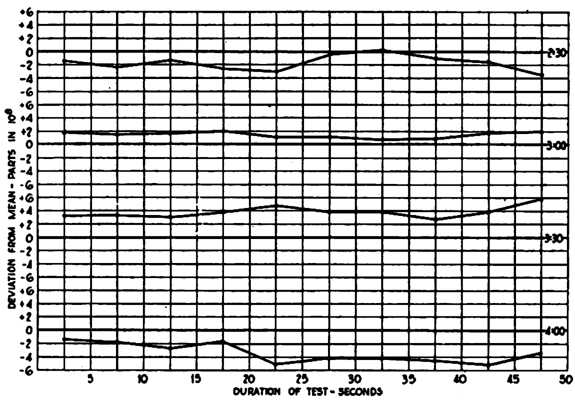

Fig. 13—A Relative Variation in Frequency of Similar Crystal-Controlled Oscillators.

this purpose a second oscillator, similar to that already described, was adjusted to approximately the same frequency, the actual difference being 0.2 cycles per second. Data taken by measuring the duration of ten consecutive periods at half-hour intervals show the relative variations. The results of these measurements are given in Fig. 13. Each point represents the difference between the ratio of the rates for any one interval from the average ratio of the rates for the entire two-hour interval. It will be noticed that for some of the runs, particularly that made at three o'clock the ratio of the rates remains practically constant. At other times, however, as at two-thirty and three-thirty, there is a marked periodicity to the variation. This may be accounted for in terms of temperature variation; the periodicity of the variation is practically the same as that of the operation of the thermostats. Whenever the two thermostats are operated out of phase, there

is a noticeable change in the ratio of the rates; whenever they are operated in phase, their rates remain practically constant.

In view of the fact that the accuracy to which a frequency may be defined depends on its constancy, on the precision of comparing its rate with the rate of the earth's rotation, and on the accuracy to which the latter rate can be determined, it seems reasonable, in view of the above data, to expect that in the near future it will be possible to define a frequency at any instant to an absolute accuracy of at least one part in ten million.

It has not been unusual to determine frequency in terms of some other quantity such as the acceleration due to gravity, the velocity of light, or the product of inductance and capacity. These are, of course, indirect measurements because each of the other quantities must first be defined in terms of time. In view of the present accuracy obtainable in the measurement of frequency, however, it is now possible to invert the procedure and measure some of these other physical quantities in terms of frequency with an accuracy heretofore unattainable.

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A RADIO-FREQUENCY OSCILLATOR FOR RECEIVER INVESTIGATIONS

By

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Summary—*A modulated radio-frequency oscillator is described. This apparatus incorporates a means for obtaining radio-frequency outputs of widely varying range, a metering system, a means for changing the generated frequency in small steps to either side of a given frequency, and a modulation and indicating system.*

A modulated radio-frequency oscillator forms a piece of apparatus which is extremely useful in an experimental laboratory. By incorporating a number of special features, such an oscillator can be made very flexible and its uses varied.

The equipment to be described was originally built in connection with work done in determining certain radio receiver characteristcs.[1] The oscillator has also been used for providing known fields in the measurement of field intensities, determining coil resistance and inductance, lining up radio circuits, and for other general laboratory purposes.

The design was such as to satisfy as far as possible the following requirements:

1. The oscillator output should be widely variable without any appreciable change in frequency.
2. A means for reading the output should be provided.
3. The frequency should be variable over a given range (such as the broadcast band) and the calibration should be maintained.
4. At any frequency setting, it should be possible to vary the frequency in small continuous steps to either side of the setting with a fair degree of accuracy.
5. Modulation should be obtainable up to at least 70 per cent with means for indicating continuously the percentage of modulation.
6. The stray field from the apparatus should be low compared to the field produced by the known radiator used.
7. Mechanical details should be such as to enable the operator to take data quickly and easily.

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[1] Proceedings of the Institute of Radio Engineers, Vol. 15, No. 5; May 1927, Page 389.

The apparatus to be described meets these requirements satisfactorily. Many details have been thoroughly developed and retained in this device, with the result that the oscillator may appear rather large and bulky. However, when it is considered to what varied use it may be put, its large size is justified.

Electrical Construction

The complete apparatus consists of a modulated oscillator-power amplifier arrangement, with an output and measuring circuit. A schematic diagram appears in Fig. 1.

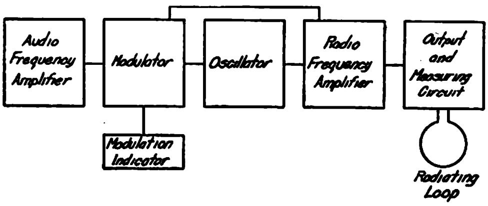

Fig. 1—Schematic Diagram of Modulated Oscillator Amplifier.

General Circuits

It was found that the master-oscillator-power amplifier method for the generation of radio-frequency currents had several advantages over the use of a simple oscillator. The oscillator employs a UX-201-A tube in an ordinary Hartley circuit. A grid condenser of 500 micro-microfarads with a grid leak of 20,000 ohms gives sufficient constancy of frequency for small variations in filament and plate voltage, without greatly reducing the radio-frequency output. For the broadcast band, an inductance of 82 microhenrys with a variable condenser of 1,000 micro-microfarads maximum capacity forms the tuned circuit. The variable condenser, which is of the straight line frequency type, is purposely made large, so that only the really straight portion of its frequency calibration is normally used, while still maintaining a three-to-one range in frequency. In order to minimize the effect of tube capacity in different tubes on the frequency calibration, a small fixed condenser of about 60 micro-microfarads capacitance is placed in parallel with the main tuning condenser. This addition was found to improve the frequency-setting relation of the main control. Fig. 2 shows this relation for the apparatus constructed.

A mechanical arrangement is employed for varying the frequency in small steps. This device is explained in detail below.

Frequencies in steps of 1,000 cycles up to 8,000 cycles to either side of a given frequency can be obtained. The calibration of this adjustment was carried out as follows:

With the vernier control at its zero position, a zero beat was produced in a radio receiver with a separate external oscillator, the frequency of which is known to be constant over the period of the calibration. When the vernier is moved off its zero position,

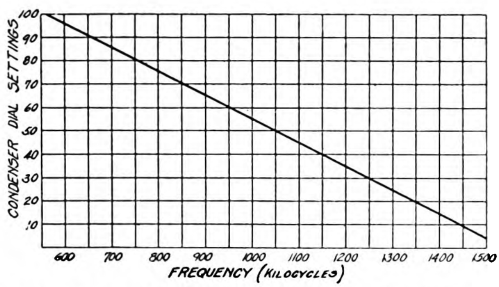

Fig. 2—Apparatus Frequency-Setting Relation.

a beat note will occur, which is compared with a calibrated audio-frequency source, adjusting the pitch of the beat-note so that it is equal to the known audio frequency. This operation was repeated for a number of frequencies up to 8,000 cycles to either side of the vernier zero position.

Since the vernier control is constructed so as to move the tuning condenser a constant amount, in angular motion, for a given rotation of the vernier, and since the variable tuning condenser has a straight line frequency characteristic, the calibration of the vernier will hold for all radio frequencies. This was checked by repeating the above beat-note comparisons for various condenser positions and the vernier calibration was found to be practically constant over the whole frequency range.

The radio-frequency power amplifier has its grid circuit coupled to the oscillator circuit by means of an untuned inductance. Two UX-112 tubes are employed in parallel and these are neutralized. The power amplifier is operated as a "proper amplifier" and its grid excitation is such as to obtain maximum output under these conditions. The amplifier plate circuit is tuned by a condenser placed on the main tuning condenser shaft.

The output circuit is coupled to the amplifier plate coil, its coupling being continuously variable for output control. It was

found desirable to have quite a large range of output—at least 5,000 to 1 in current. The combination of a radio-frequency current transformer, thermo-couple milliammeter and switching arrangement has been found satisfactory for providing and metering such a widely varying output.

Current Transformer for Provision of Various Output Currents

The current transformer for radio frequencies,[2] in the form employed in this apparatus, has an iron core of 10 mil (0.0254

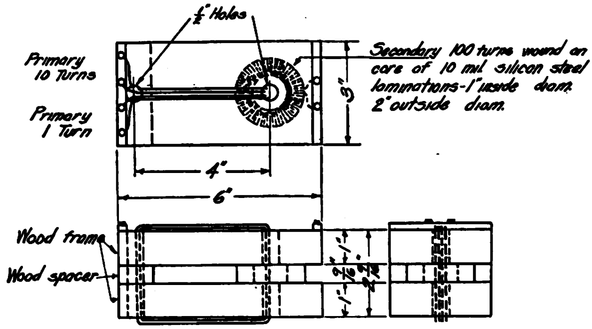

Fig. 3—Current Transformer for the Measurement of Output Currents.

centimeters) toroidal silicon steel laminations, $^9/_{16}$ inch (1.43 centimeters) thick, 2 inches (5.08 centimeters) outside diameter, and 1 inch (2.54 centimeters) inside diameter. The core is covered with insulating oiled silk and one winding of 100 turns is evenly spaced over this core. The other winding of either 1 or 10 turns is looped through the hole in the toroid. Wood blocks keep the windings well spaced from one another so as to reduce capacity effects between the windings and maintain the current ratio. The construction is shown in Fig. 3.

It can be shown that the current ratio for such a transformer is given by an expression of the form

$$\frac{I_1}{I_2}=\frac{n_2}{n_1}\left(1+\frac{\alpha R_2}{\omega L_2}\right)$$

[2]"Measurement of Alternating Electric Currents of High Frequency"—D. W. Dye, *Proceedings* of Royal Society (of Great Britain) A-1914, Vol. 90, Page 621.

Bureau of Standards Circular No. 74, Page 153.

"Current Transformer Methods of Producing Small Voltages and Currents at Radio Frequencies for Calibrating Purposes"—D. W. Dye *Journal* Institution of Electrical Engineers (London) June 1925; Vol. 63, Page 597.

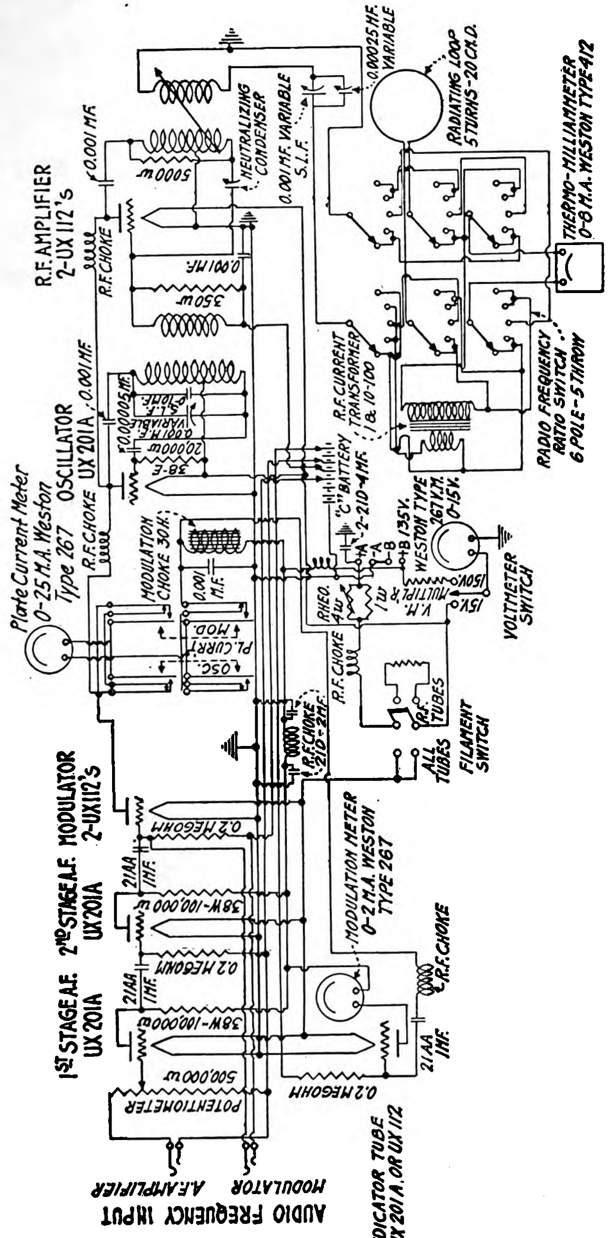

Fig. 4—Complete Circuit Diagram of the Apparatus.

where I_1 and n_1 are the primary current and turns number respectively.

I_2 and n_2 are the secondary current and turns number respectively.

R_2 and L_2 are the secondary resistance and inductance respectively.

α is a coefficient depending on the iron losses, having a value usually less than 1.

When $\frac{\alpha R_2}{\omega L_2}$ is small compared to 1, this expression reduces to

$$\frac{I_1}{I_2}=\frac{n_2}{n_1}$$

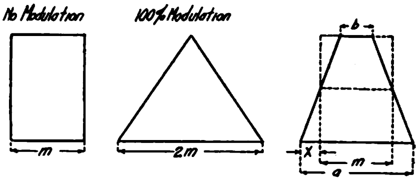

Fig. 5—Cathode-Ray Oscillograph Figures Used in Modulation Percentage Determination.

The secondary circuit contains the ammeter resistance and, for accuracy of transformation, this must be made low compared to the inductive reactance of the transformer secondary. For the transformer of the dimensions given used at frequencies in the broadcast spectrum, a 43-ohm thermo-milliammeter (0–8 milliamperes) has been found satisfactory. The ratio was checked by measuring the currents in both windings and an accuracy of better than 5 per cent was found throughout the range of currents and frequencies required.

Fig. 4 shows the complete wiring diagram of the apparatus and from it may be seen the method of using the current transformer. A six-pole, five-point switch allows the meter to be used in either the primary or secondary of the current transformer so that current ratios of 1–100, 1–10, 1–1, 10–1, or 100–1 may be obtained between the meter circuit and the load or radiating loop circuit. The capacity between various switch parts has been kept low. In addition, the effects due to any undesirable capacities are minimized by the use of low impedance load and metering circuits.

Provision of Known Fields

The radiating loop with a known current in it may be used to induce an easily calculated voltage in a second loop having a known mutual inductance between loops. Known field intensities in a plane x centimeters from the loop and parallel to the plane of the loop can be calculated from the field H which is

$$H = \frac{2\pi a^2 in}{10} \times \frac{1}{(x^2+a^2)^{3/2}} \text{ (gausses)}$$

The field intensity in microvolts per meter is $E = 300H \times 10^8$. Thus,

$$E = 18850 \frac{na^2 i}{(x^2+a^2)^{3/2}} \times 10^6 \text{ microvolts per meter}$$

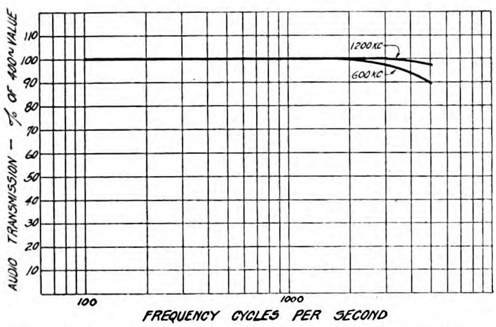

Fig. 6—Overall Audio-frequency Characteristics of the Apparatus.

where a is the radius of the loop in centimeters, i is the loop current in amperes, n is the number of turns on the loop.

If the plane in which it is desired to know E is at an angle to the plane of the loop, the field intensity may be calculated from the above expression, introducing a factor of the cosine of the angle between planes. Field intensities ranging from ten to one million microvolts per meter are readily obtained in this manner.

Modulation

The conventional plate-circuit choke coil method is used for modulation, two UX-112 tubes controlling both master-oscillator and power amplifier. This system has been found satisfactory for the purpose. Fairly high degrees of modulation are obtainable. The relation between input voltage and percentage modulation has been found linear up to 70 per cent. For measuring the per-

centage of modulation, the ordinary method of measuring the audio-frequency voltage across the modulation reactor and comparing this value with the direct current plate voltage of the oscillator and modulator will not apply. This is due to the fact that both the oscillator and radio-frequency amplifier are being modulated. The cathode ray oscillograph has been used for modulation-percentage determination. One set of the oscillograph plates is excited from the voltage induced in a pick-up coil coupled to the radiating loop. The oscillator is modulated from an audio-frequency source. The other set of oscillograph plates is supplied with voltage from the same audio source. In general, the resulting oscillograph figure is a trapezoid. The various forms it may take are shown in Fig. 5.

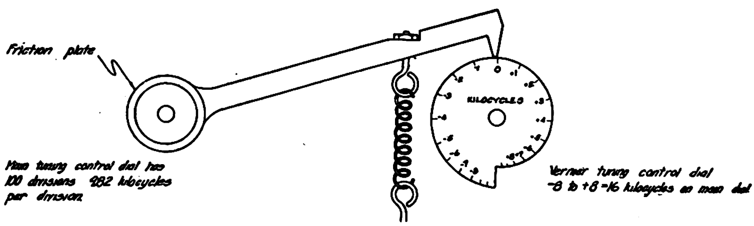

Fig. 7—Cam and Lever Used for Vernier Adjustment.

Using the symbols of Fig. 5, the modulation is given by $\frac{2x}{m} \times 100$ per cent. But $m = \frac{a+b}{2}$ and $x = \frac{a-b}{4}$, therefore modulation

$$m = \frac{a-b}{a+b} \times 100 \text{ per cent.}$$

As an indicating device, a UX-201A tube is employed with its normal plate current biased to practically zero and its grid excited from the audio voltage developed across the modulation reactor. The plate circuit of this tube contains a 0–2 milliampere meter, the scale of which has been calibrated directly in modulation percentages using the oscillograph method described. A radio-frequency choke and by-pass condenser prevent the radio-frequency voltage from affecting the indicator tube reading. With the radio-frequency amplifier tubes operating properly and the circuits equalized, the modulation indication is practically constant with varying radio frequency.

A two-stage resistance-coupled amplifier supplies sufficient voltage to the modulator grids. Its characteristic is practically flat over a wide range of audio frequencies.

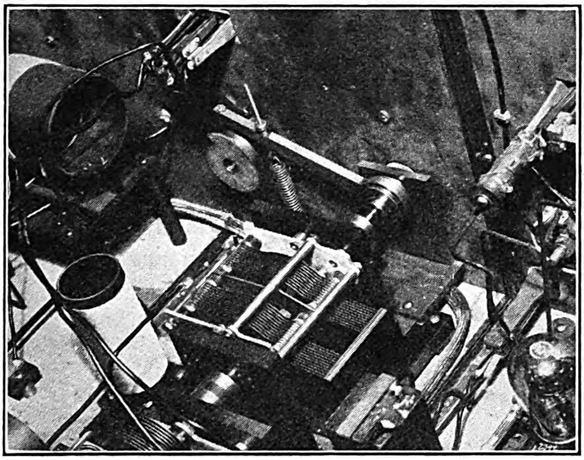

Fig. 8—View of Cam and Main Tuning Condenser

Equalization of Output Current

The radio-frequency output current was found to vary considerably with frequency and a constant output was believed to be desirable. A means for correcting this fault to a considerable extent was devised. This consisted of placing suitable fixed

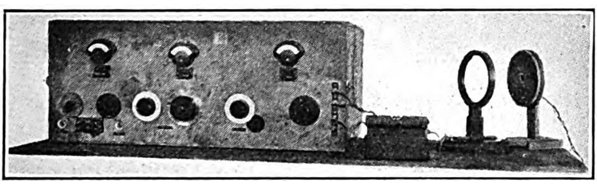

Fig. 9—External View of the Complete Apparatus.

resistances across both the radio-frequency amplifier coupling coil and its output coil. The variation of output current in the radiating loop was thus made less than 15 per cent for a constant output coupling setting over the frequency range. These resistances were also found to lessen the effect of the tuned circuit on the audio characteristics.

Overall audio-frequency characteristics from audio-amplifier input to modulated radio-frequency output are shown in Fig. 6. The data for these curves were taken by rectifying some of the voltage picked up with a biased detector tube and observing the audio-frequency output. A slight effect of resonant circuit selection still remains, but this is not of any considerable magnitude and can be allowed for in exact measurements.

Mechanical Construction

Tuning and Vernier Controls

The oscillator and output tuning condensers are both mounted on the same shaft and are controlled from a single dial on the front

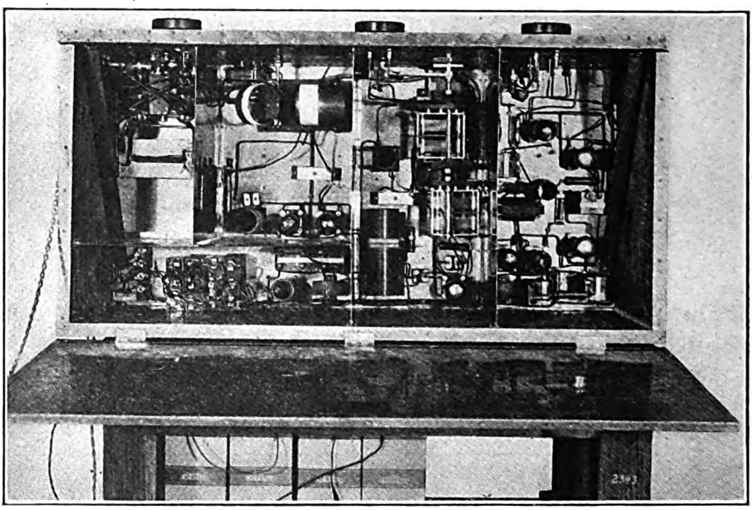

Fig. 10—Internal View of the Complete Apparatus.

panel. A friction cam arrangement has been found suitable for vernier adjustment. This consists of a lever and friction plate on the main shaft and a cam on a separate vernier shaft. The end of the lever arm bears on the cam and is held in position by a strong spring. Any movement of the vernier shaft will cause the lever to move either up or down a small amount and, since the friction plate is now bound to the main shaft, the main tuning condensers will move a small amount. The main tuning shaft can be moved without disturbing the vernier arrangement, since the friction plate will slip and the vernier lever arm will remain fixed due to

the combined action of the spring and the cam rest. The criterion of whether such a system will operate is that the force required to move the friction plate over its bearing should be greater than the bearing friction in the tuning condensers. In practice, this is easily accomplished. The details may be seen in Figs. 7 and 8.

The whole apparatus is mounted in a double copper shielded box. The front panel and base board have been made removable as a unit for convenience. The various units, such as audio-frequency amplifier, oscillator, radio-frequency amplifier, and output circuit, are shielded from one another. The details may be seen in Figs. 9 and 10. The grid batteries are located in a separate compartment, but the filament battery of six volts and the plate battery of 135 volts are located outside the shielded box. Radio-frequency filters in the battery leads prevent radiation from these.

The thermo-milliammeter for the measurement of output is placed externally near the apparatus, a twisted pair of leads of short length being used for connections.

In conclusion, we desire to mention that this design was carried out under the direction of Mr. Julius Weinberger, whose suggestions were very valuable in the development of the equipment.

ON THE INFLUENCE OF SOLAR ACTIVITY ON RADIO TRANSMISSION*

By

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(Laboratory for Special Radio Transmission Research, Bureau of Standards)

Summary—The paper describes further examination of the daylight long wave signal measurements of the Bureau of Standards for evidence of correlation with solar activity. It has already been shown that a probable correlation exists between signals and sunspots when the observations are continued for several years and are averaged in periods of a month or more. The present paper deals especially with observations averaged in shorter (5-day) periods. Here while the relationship is generally evident it is sometimes obscured by an apparent relative phase shift between the signal and sunspot curves.

At the meeting of the American Section of the U.R.S.I. in April, 1927, some curves were shown[1] indicating a relationship between very long wave transatlantic radio reception and solar activity as measured by sunspot numbers and by the values of the solar constant. These curves represented annual averages from 1915 to 1926, and monthly averages from 1922 to 1926. These indicated without much question that there exists a direct correlation between solar activity and daylight signals, when averaged over long periods. Curves were also shown illustrating the correspondence of 27-day periodic averages of sunspots and reception.

G. W. Pickard[2] had previously published a paper using his own observations in the broadcasting band, as well as the observations of others, on very long as well as ultra short wavelengths, in which he produced striking evidence of the dependence of radio phenomena on solar activity. Mr. Pickard's own observations indicated quite positively an inverse relationship between the strength of night signals in the broadcasting band and sunspot numbers, while the ultra short wave night signal material treated by the same methods, indicated a direct correlation. This last conclusion is, however, seemingly not in agreement with a number

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Read at the International meeting of the Union of Scientific Radio Telegraphy in Washington, October 14, 1927.

[1] Proc. I. R. E. Vol. 15, p. 825, 1927.

[2] Proc. I. R. E. Vol. 15, p. 83, 1927. See also Vol. 15, p. 749, 1927.

of observations made by others of weakened ultra short wave night signals at times of great solar and magnetic disturbances.

During the past summer the examination of the long wave observational material at the Bureau of Standards, in connection with sun spots, has been continued. Especial attention is now being given to the correlation of the daily observations, or the averages of a few days at most, rather than to long time averages. It is found in general that daily comparisons of signals and sun spots give too complex curves to be of use. But when they are

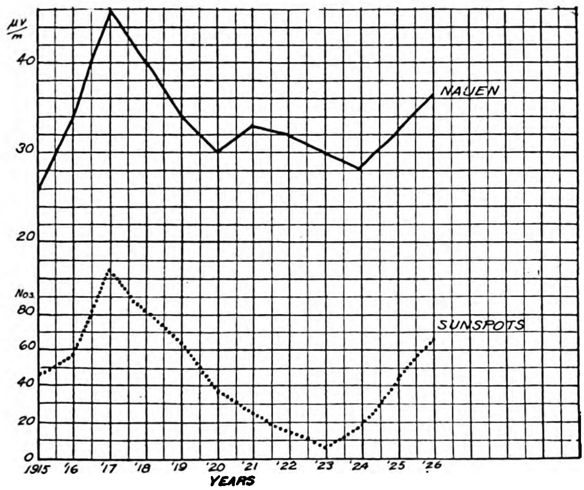

Fig. 1—Annual Average Signal Intensity of Nauen (AGS), 10 A.M. and Sunspot Numbers.

averaged in five-day periods good results begin to appear. It is also found in the examination of the material that the stations at moderate distances (less than 1000 km.) when treated in this way give on the whole better correlation with sun spots than more distant stations.

Since the signal and sunspot curves do not follow each other entirely in detail, it seems probable that there are other factors besides solar activity which contribute to the signal variations. One of these we believe to be temperature which has already been shown[3] to have a strong influence especially in winter on the reception in Washington of the transatlantic stations at New Brunswick and Tuckerton, New Jersey.

[3] Proc. I. R. E. Vol. 14, p. 781, 1926.

There also seems to be a tendency for the sunspot and signal curves to shift relatively back and forth in phase as though at times there might be a lag in the appearance of the sun spots in respect to the fundamental solar disturbance producing them, while on the other hand, there often seems to be a delay in the formation of the conditions in the earth's atmosphere which produce the changes in signal strength. At any rate there is no doubt that at times the increase in signal precedes the increase in sun spots, while at other times the reverse seems to be true.

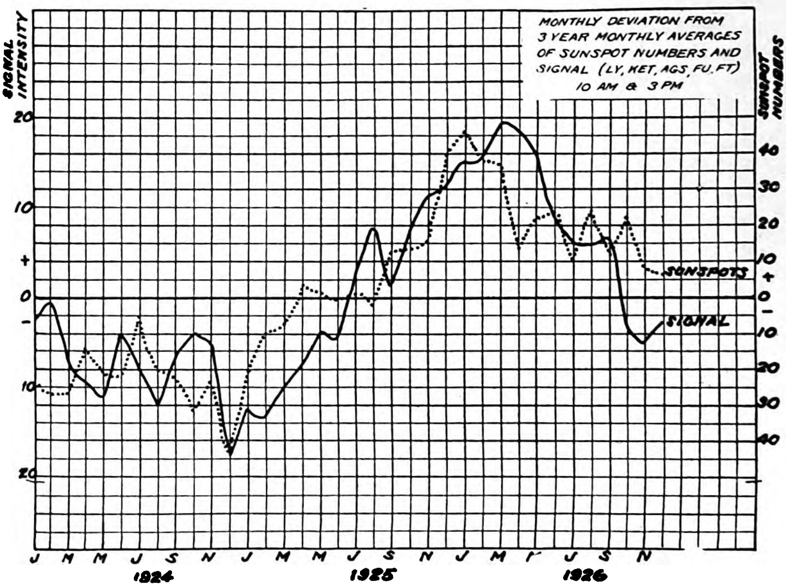

Fig. 2

Notwithstanding these phase shifts, the frequent similarity of the sunspot and signal curves renders it almost certain that we are dealing with a genuine relationship. On the whole there seems to be about the same degree of correlation between the signals and sun spots as between signals of similar wavelengths received at a given receiving station from two separate transmitting stations. In the following, Figs. 1, 2, and 3 deal with long time comparisons in which the connections between solar activity and radio transmission is believed to be fairly well established, while Figs. 4, 5, 6, and 7 deal with short period (5-day) averages in which the transmission curves are smoothed by the smoothing formula

$\frac{a+2b+c}{4}$. This last, without changing the positions of the main maxima and minima, renders the transmission curves more easily comparable with those of the sun spots.

Fig. 1 shows the annual averages of Nauen ($\lambda = 13000$ m.) as received at Washington at 10 A.M. from 1915 to 1926, and the corresponding annual averages of the Wolfer sunspot numbers covering the same period. The signals from 1915 to 1921 were taken by the shunted telephone method and are therefore less

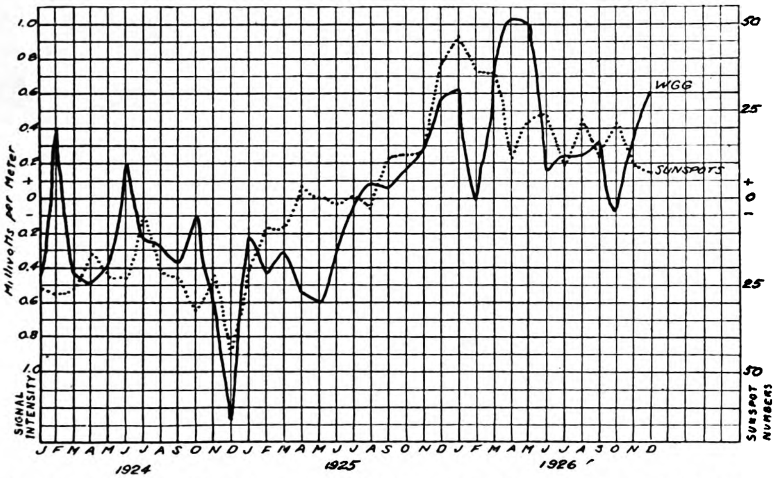

Fig. 3—Monthly Deviation from 3-Year Monthly Averages of Tuckerton (WGG), 10 A.M. and 3 P.M. (Corrected for Variations in Antenna Current) and Sunspot Numbers, 1924–1926.

accurate than those taken later, but we believe that there can be no doubt that the general course of the curve is correct.

Fig. 2[4] shows the deviations of the individual monthly averages from the three-year monthly averages of five distant stations as compared with the sunspot numbers. It is seen that, while the two curves do not follow each other in detail, in general they rise together during the three years of increasing solar activity. The portion of the figure dealing with the rapid increase of sun spots in the autumn of 1925 is of especial interest.

Fig. 3 shows similar curves for the Radio Corporation of America transatlantic station at Tuckerton, N. J. (WGG), only 250 km. from Washington.

[4] Figs. 1 and 2 are repeated from PROC. I. R. E., Vol. 15, p. 825, 1927.

As has been said, the degree of correlation between sun spots and signals is comparable with that between two signals received

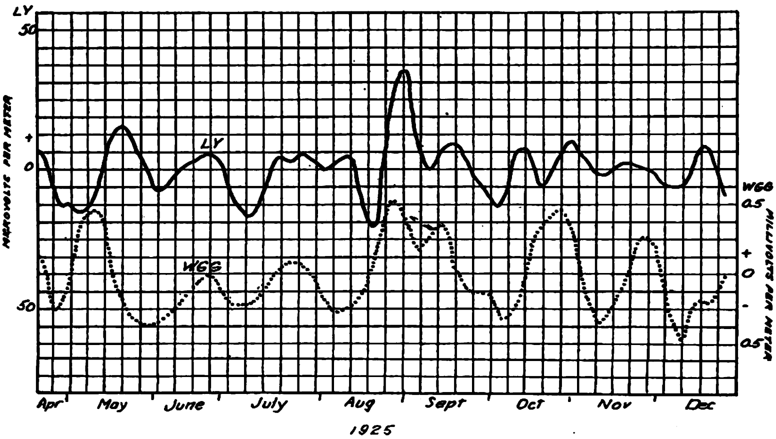

Fig. 4—Deviation of 5-Day Averages from Monthly Averages of Tuckerton (WGG), 10 A.M. and 3 P.M. and Lafayette (LY), 10 A.M. (1925).

at one point from different transmitting stations. For comparison with the following curves, therefore, Fig. 4 is shown giving the

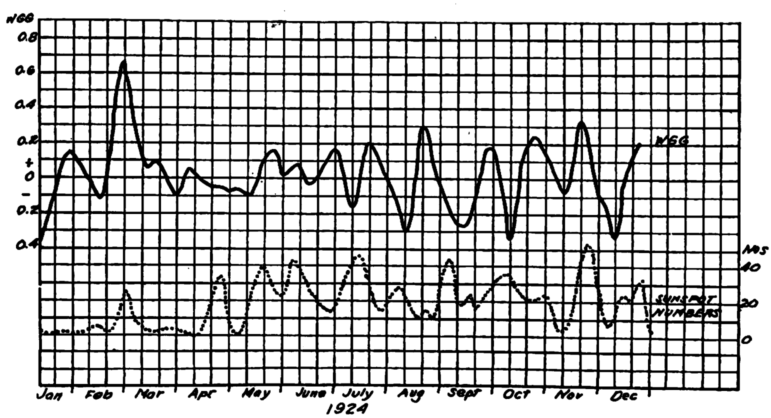

Fig. 5—Deviation of 5-Day Averages from Monthly Averages of Tuckerton (WGG), 10 A.M. and 3 P.M. and 5-Day Averages of Sunspot Numbers, 1924.

relationship of the five-day averages of Tuckerton, WGG, and Bordeaux LY as received at 10 A.M. during a portion of 1925 at the Bureau of Standards.

Fig. 5 shows the deviations of the five-day averages of Tuckerton as compared with the five-day averages of the sun spots in 1924. The year 1924 was very near the sunspot minimum.

Fig. 6 shows the relationship between the sun spots and the same station between March, 1926 and March, 1927. Here the curves are drawn on a smaller scale than in 1924 on account of the much greater amplitude of the variations both of the sunspot numbers and of the signal intensity. It is at once seen that the degree of correlation of the sunspot numbers and signal seems much better in 1926 and 1927 than during the period of low solar activity in 1924.

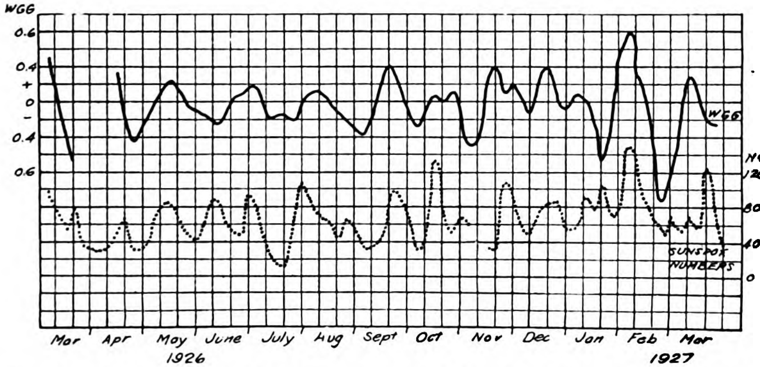

Fig. 6—Deviation of 5-Day Averages from Monthly Averages of Tuckerton (WGG) and 5-Day Averages of Sunspot Numbers, 1926–1927.

Fig. 7 shows similar curves for the R. C. A. station WSS at Rocky Point, L. I., at a distance of 435 km. from Washington in 1926–27. In addition to the signal and sunspot curves, a curve of the five-day averages of the range of the horizontal intensity of the earth's field as observed at Cheltenham, Maryland, is included. The resemblances between these curves are really striking, though we find here also the relative shifts in phase already mentioned between the sunspot numbers and the signal intensity. In some cases the signals evidently lead, while in other cases they lag.

In examining the curves of the five-day averages, it is evident that while there is little question of a connection between solar activity and radio signal intensity, yet there is a lack of any direct proportionality between the sunspot numbers and signal strength. For it is not by any means always the largest increase in sun spots

which corresponds to the largest increase in radio transmission.

Our present knowledge concerning the relation between solar activity and the strength of radio transmission may be briefly summarized as follows:

It seems reasonably certain that long wave daylight signal strength has increased in recent years with the increasing solar activity, when averaged in periods of a month or more, and there is fair evidence that the annual averages of signal intensity have roughly followed the sunspot curve since 1915.

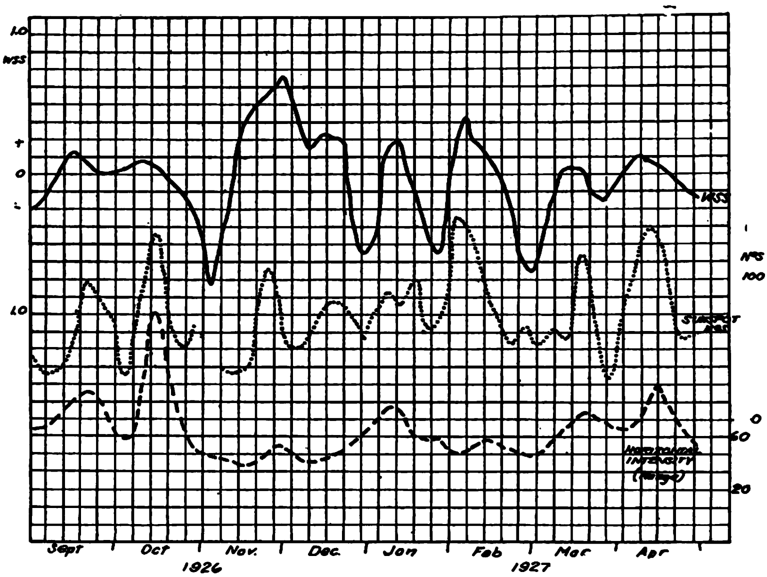

Fig. 7—Deviation of 5-Day Averages from Monthly Averages of Rocky Point (WSS), 10 A.M. and 3 P.M. and 5-Day Averages of Sunspot Numbers and Horizontal Intensity (Cheltenham), 1926–1927.

It is also reasonably certain that night transmission in the broadcasting range when averaged in periods of a few days is reduced at times of high sunspot numbers.

The information regarding the effects of the sun on long wave daylight transmission and ultra short wave night transmission averaged in periods of a few days, is somewhat discordant differing at different times and on different stations. These discrepancies may be due to the relative phase shifts in the transmission and sunspot curves already mentioned.

Very few observations have been published on long wave night transmission, on broadcasting day transmission, nor, so far as is known, have any been made on broadcasting night transmission extending over more than one year.

There is in general a rough resemblance observed in the reception curves of different long wave stations lying in different directions from Washington. This would seem to imply either that there are simultaneous changes in the electrical conditions in the upper atmosphere over very large areas or that the conditions in the immediate neighborhood of the receiving station form the chief controlling factor in signal strength. Occasionally, however, marked differences in reception from stations in different directions have been noted, a phenomenon which is much more common in the case of the broadcasting and ultra short waves at night.

While the sunspot numbers give a rough indication of the conditions governing transmission, it seems probable that they are only an imperfect index of the particular form of solar activity which apparently is the chief agent in controlling radio phenomena.

IONIZATION IN THE UPPER ATMOSPHERE*

By

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THE more important agencies which may conceivably cause the ionization of the upper atmosphere of the earth are the ultra-violet light, α and β particles, all of solar origin, the penetrating radiation of cosmic origin, and the ionizing radiations from terrestrial sources. The last mentioned may perhaps be ruled out immediately because of the fact that the conductivity of the lower atmospheric strata increases rapidly with the height for the first few kilometers. The possible effects of these ionizing agencies have been considered in papers by Chapman and Milne, Benndorf, Elias, Lassen and others. Recently experiments with the electromagnetic waves of radio telegraphy together with theories of the propagation of these waves over the surface of the earth have led to information about the ionization in the upper atmosphere more definite than hitherto obtainable, and it has been of interest to examine again the causes of the ionization.

The experiments of Breit and Tuve with 70-meter waves, of Appleton with 400-meter waves, of Heising with 57- and 111-meter waves, of Wagner and Quack with 15- and 16-meter waves, of Hollingsworth and Eckersley with long waves, and the experiments of Taylor on the skip distances of waves below 50 meters and the theoretical considerations of Taylor and Hulburt on these skip distances, show that the electron density N increases with the height Z above the earth reaching a value of about 4×10^5; above the height where N has this value the electron density is not known except that it does not go on increasing. Although the radio data are none too extensive, it may be taken that in the day time (for the North Temperate Zone) the height where N is of the order 10^5 is roughly 150 to 200 km.

Because of the diurnal variation in the ionization, we choose the ultra-violet light of the sun as being the ionizing agency deserving first consideration. In order to make an explicit calculation

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of the ionization the temperature, the pressure, and the constituent gases and their partial pressures must be known at each height in the upper atmosphere. These we may take as given completely in the classical thermodynamic isothermal equilibrium theory of Humphreys, Jeans, and others. There is a question as to the existence of hydrogen in the upper atmosphere, but the conclusions given later are much the same whether hydrogen is there or not. There is also the question of ozone, or oxygen, which may be of great importance. The law of the recombination of the electrons with the positive ions must be known. For this we have J. J. Thomson's theory of recombination, and complete formulas are available. We must also recognize the possibility of the electron attaching itself to a neutral molecule, for when it does this, thereby producing a negative ion, it is no longer an energetic refractor of the radio waves. The oxygen molecule is the only important one in this connection, and the values of the attachment coefficient measured in the laboratory for pressures of 10 mm. of mercury and above must be extrapolated to pressures below 10^{-2} mm., perhaps a questionable extrapolation. The diffusion of the electrons and ions must be considered; this is a very important influence. Using all these things and making entirely acceptable assumptions as to the amount of ultra-violet light from the sun in the spectral region useful for ionization the N,Z curve rises from $N \doteq 0$ at 50 km. to the order of 10^5 at $Z = 140$ km. in the winter time; above this height N falls off rapidly. In the summer time the electron bank is higher and denser than in the winter. The N,Z curve is in fair accord with the radio data for full daylight conditions, and its change at night is, as far as can be seen, in agreement with night time radio conditions. Below the electron bank there is an ion bank. The exact density of the ions is uncertain, but it seems possible that these may persist in sufficient quantities to play a part in the refraction of the longer radio waves. The uncertainty is due to the lack of exact knowledge of the formation and recombination of the positive and negative ions; the question seems to be inextricably bound up with and confused by the presence of ozone, the laws of ozone formation and destruction being quite unknown.

On the whole it seems that the ultra-violet light of the sun is a necessary and sufficient cause of the Kennelly-Heaviside layer. This means that it is not necessary to consider possible effects of other agencies of ionization, such as α and β particles etc.,

except as secondary, or unusual effects, which may of course be quite important, as the recent correlations between sun spots and radio transmission indicate. In conclusion we must emphasize the view that we can see no possibility of the existence of electron banks above the main one which has its maximum electronic density at a height around 150 to 200 km. and therefore that inferences from radio data which suggest the presence of such outlying layers must be examined with care before they can be accepted.

A THEORY OF THE UPPER ATMOSPHERE AND METEORS*

By

H. B. Maris

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Summary—*A calculation of the rate of separation of gases of different density in the earth's atmosphere leads us to expect a uniform mixture of all gases below 100 km and densities of hydrogen and helium roughly a hundred thousandth of those previously calculated, for greater heights. Known absorption and radiation coefficients for gases of the upper atmosphere indicate that we should expect a daily temperature variation of about 140° during the summer and 30° during the winter for all heights greater than 80 km. Carbon dioxide is found to be more effective than water vapor in determining the final escape of radiation from the earth and the conclusion is drawn that variations in the carbon dioxide content of the air may explain the variation in climatic conditions of the earth indicated by the ice ages of the past. Frictional resistance offered by the upper atmosphere to the passage of meteors through it is not sufficient to account for the energy radiated by meteors and the conclusion is reached that the energy of the meteor is probably dissipated into the air by the escape of atoms and molecules driven from the meteor by the energies of impacts with molecules of the air.*

THE force of gravity acting on the atmosphere of the earth causes the heavier gases to settle downward and the lighter gases to rise to higher altitudes by diffusion but winds unhindered by diffusion would by convection keep the composition of the air uniform at all elevations. The classical ideas of the change in atmospheric pressure with altitude (e.g., Humphrey, Jeans, Chapman and Milne, etc.) have been based on the assumption that convection is negligible, at least in the upper atmosphere, and that each gas is, through diffusion, in gravity equilibrium with its own partial pressure. Investigation has shown, however, that diffusion is of importance only at elevations greater than 100 km.

The ordinary equations of diffusion show at once that if the air were uniformly mixed at all altitudes and then left free from all convection currents, there would be a constant flow of lighter molecules upward and of heavier molecules downward, which would be independent of the altitude until a level was reached

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where the diffusing gas would be in gravity equilibrium. This *diffusion* level for hydrogen would move from infinity down to 142 km. in one day, at the end of five days it would be at a height of 127 km. and in 50 days it would be at 113 km. The corresponding levels for helium would be at 137, 120 and 106 km., respectively. The new calculations give hydrogen and helium densities above 150 km. roughly 1/100,000 of the values previously calculated.

Recent use of the upper atmosphere as a medium for transmitting electromagnetic radiation of wavelength 10 to 10,000 meters has emphasized the importance of knowledge, or at least of a theory, of changes which occur in the upper atmosphere between day and night conditions during different seasons of the year. Absorption of solar and terrestrial radiation by the air must determine any theory of temperature distribution in the upper atmosphere. Humphrey has discussed this problem and has suggested that it should have been worked out, but no attempt has been made previously to apply radiation and absorption coefficients and solve for the thermal condition of the upper atmosphere or to estimate probable temperatures at elevations greater than 20 km. for the radiation conditions of day and night or winter and summer.

Water vapor above 11 km. absorbs a little over 20 per cent of black body radiation from below at earth temperatures while carbon dioxide absorbs nearly 40 per cent. Ozone absorbs only about 2 per cent but its presence is important because it absorbs about 4 per cent of the solar radiation at an altitude where most of the re-radiation must be by the ozone itself. Temperature calculations based on these absorption coefficients show that for a 50 deg. latitude above a height of sixty kms. we should expect a temperature of about 250 deg. K during a winter day with a drop to 220 deg. during the night and a temperature of 370 deg. during a summer day with a corresponding drop to 230 deg. during the night. The atmosphere at the base of the stratosphere cannot be in radiation equilibrium, since its temperature varies from 205 deg. K over the equator to 230 deg. K over the poles but must receive more radiant energy than it loses both from above and below during a 24-hour day. The temperature condition of the earth's surface is in very unstable equilibrium. The loss in heat by radiation from the warm equator is much less than from the cooler polar regions. An increase in temperature at sea level near the equator would not result in an increase in the energy lost by radiation from these regions,

but would actually result in a decrease. Loss of heat by radiation from the earth depends, not on the condition of the surface, but on the temperature at the base of the stratosphere and absorption in the stratosphere. A slight change in the carbon dioxide of the air would have a tremendous influence on the climate of the earth. If the carbon dioxide content of the air were increased from the present 0.03 per cent to 0.1 per cent tropical plants would probably grow in the polar regions. On the other hand, if this protecting sheet decreased from 0.03 per cent to 0.01 per cent, ice would probably be found near the equator.

Since the present theory leads to low densities of the atmosphere above heights of 100 km., and densities one hundred thousandth of those of classical tables at 300 km. the facts about the appearances of meteors require explanation. It seems possible to do this following to a certain extent the ideas of Sparrow and to a certain extent those of Lindemann. When a high speed meteor strikes an air molecule, it is assumed that the energy of the impact violently ejects atoms, molecules and possibly small particles of molecular dimension from the body of the meteor. This ejected material by virtue of its velocity carries into the air the energy which eventually gives the light of the meteor trail. For example, when a nitrogen molecule strikes an iron meteor which has a velocity of 40 km. per second, the energy of the impact is sufficient to raise the temperature of 1800 molecules 1000 deg. C or to evaporate 56 molecules of iron, or to evaporate and ionize 24 molecules of iron. As a result of this impact a mass many times that of the nitrogen molecule is ejected from the meteor principally in the form of highly energized iron atoms which have velocities slightly greater than that of the meteor itself. The inelastic collisions of these iron atoms with the molecules of the air result in the visible trail. The excitation energy of these collisions may be as high as 155 volts for nitrogen or 280 volts for argon. Much of this energy may be radiated in the ultra-violet or even soft x-ray region, and it is probable that not more than one-tenth of the total radiation is in the visible part of the spectrum. Therefore, the total mass of the meteor must be much more than that derived by Lindemann and Dobson from their considerations of the relation between the mass of a meteor and its light. The temperature changes in the upper atmosphere from evening to morning, and from winter to summer, given by the present theory lead one to expect appearance of meteors at heights which are greater by perhaps 5 km.

in the evening than in the morning and in the summer than in the winter. It would be interesting to know whether this difference has been observed.

Recent studies of the propagation of electromagnetic waves over the earth's surface have emphasized the need of a theory and definite conclusions concerning diurnal and seasonal changes in temperature and composition of the atmosphere at heights greater than 50 km. It is the purpose of this discussion to take what steps are possible toward the meeting of this need.

A RADIO FIELD STRENGTH SURVEY OF PHILADELPHIA

By

Knox McIlwain and W. S. Thompson

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Summary—*Measurements of the strength of the electromagnetic waves broadcast by station WFI were taken at representative locations throughout the city of Philadelphia and its environs. In the method used, measurements of the antenna loop resistance were used as a guide in determining when interference was sufficient to cause the rejection of a particular field strength reading. Lines of equal field strength were plotted on a map of the territory; these showed the location of several shadows and of one area in which the field strength was considerably higher than in the surrounding neighborhood. Conditions throughout the city remained virtually unchanged during the two years of the survey, with the exception of a considerable new shadow cast by a large building recently erected near the broadcasting station.*

MEASUREMENTS of the electromagnetic field strength due to a particular radio transmitting station at a definite time and place are of importance in the study of the transmission of radio waves through space. At a point far from the transmitting station the field strength varies considerably with the period of the day (or night), the season of the year, and atmospheric conditions. Studies of such variations have been conducted by a number of agencies and now include the results of several years' work. Near the transmitting station the measurements do not vary greatly with time and atmospheric conditions, but within this area, where the field strength from the given station is comparatively great, there are great differences in the intensity of the electromagnetic field at different points. These variations were naturally suspected of being caused by the topography of the surrounding country, and by the presence of metallic materials in the immediate vicinity.

A study of the conditions existing in Washington, D. C., and in New York made in 1923[1] indicated the way hills, large building areas, rivers, and other local conditions affected field strength. It was considered worth while to extend these studies to other localities, particularly large cities, where the variations were found to be most marked. Philadelphia offered local conditions differing

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[1] "Distribution of Radio Waves from Broadcasting Stations over City Districts," Bown and Gillett, I. R. E. Proc., Aug. 1924.

considerably from either of the cities previously studied. During the past two years the authors have been making measurements of the field strength produced by the transmitting station WFI in various parts of the Philadelphia area. Due to the congested condition of the air in the broadcasting frequency range, measurements were possible only during certain restricted periods each day. Those in charge of the station cooperated by carefully maintaining a constant power output and frequency.

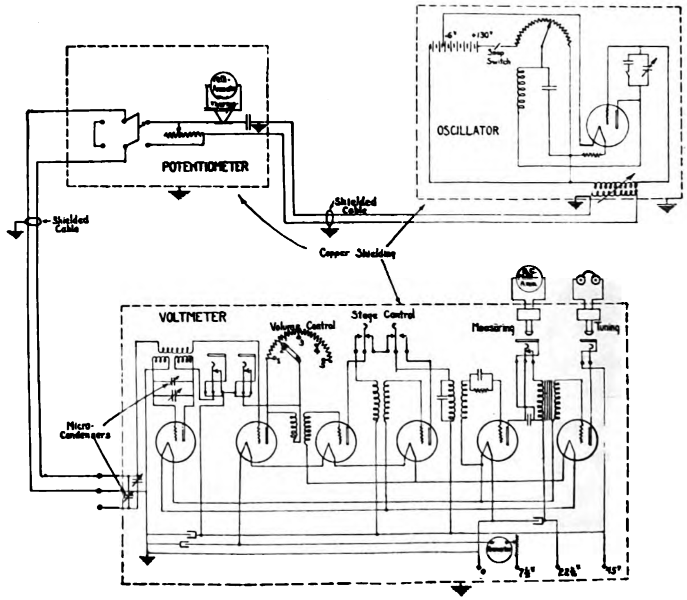

Fig. 1—Setup of Apparatus for Calibrating a Radio Receiving Set as a Voltmeter.

During the two years of the study certain readings were periodically rechecked; these readings proved to be practically constant throughout the investigation. Near the end of the study an entire recheck of the area was made. Conditions throughout the city and its environs proved to be unchanged with one very interesting exception. This exception, which showed the effect of a new steel skeleton building near the transmitting station, will be fully treated in the discussion of results.

Method and Apparatus

The method used in measuring the field strength was substantially the same as that described by Bown, Englund, and

Friis.[2] In the methods used by these investigators each reading was obtained by noting the change in direct-plate current of a detector tube caused by an incoming carrier wave. The incoming carrier was then eliminated, and a local oscillator used to impress a known variable voltage on the set. By adjusting the local voltage to cause the same variation in detector-plate current as given by the incoming carrier, the voltage, and by calculation the vertical component of the field intensity of the incoming signal was obtained. The voltage impressed across the input of a radio receiving

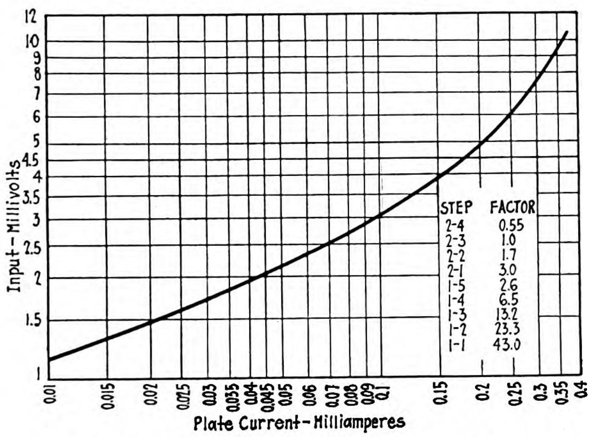

Fig. 2—Calibrating Curve of Receiving Set.

set by a given electromagnetic field depends on the effective height of the antenna used. The units of field intensity are volts, or fractions thereof, per meter.

The main difference in the method employed in the present investigation was that a set, consisting of a standard six tube superheterodyne receiver and loop antenna, was calibrated as a voltmeter in the laboratory before being used in the field (Fig. 1). The change in direct-plate current of the second detector tube due to a known voltage of a local oscillator was noted, and curves were plotted of plate current reading against impressed voltage (Fig. 2). Many calibrations were made at first to insure the reliability of the curves. It was soon found that the calibration was quite stable, changing only as the detector tube and plate batteries aged. From then on occasional check calibrations were made.

[2] "Radio Transmission Measurements," Bown, Englund, and Friis, I. R. E. Proc., April 1923.

In the calculation of the field intensity required to produce a given voltage on the input terminals of the set, the resistance of the loop circuit was required. This was obtained by the well-known reactance variation method of measuring circuit resistance, the calibrated set being used to indicate current differences due to the change in loop circuit reactance. The reactance change was caused by a calibrated micro-condenser connected permanently in parallel with the tuning condenser. In the field this micro-condenser permitted very accurate tuning. The method of measuring loop resistance was convenient in that it was very simple and capable of operation with no change in the setup of the apparatus; a fact which was utilized in determining when interference was present.

The reactance variation method of measuring circuit resistance depends on obtaining the ratio of the current in the circuit (or in this case the voltage across a particular fixed portion) at resonance, to the current in the circuit when a known reactance is introduced. When electromagnetic energy from sources other than the transmitting station is present, it will in general be made up of fields of other frequencies than that of the measured signal. In such a case it is impossible to adjust the loop circuit to resonance for all frequencies present, with the result that a greater reactance variation will be necessary to produce the desired ratio of currents. The loop circuit resistance is calculated from the following equation:

$$R=\frac{C_2-C_1}{4\pi f C_1 C_2}\sqrt{\frac{I_1^2}{I_R^2-I_1^2}}$$

where f is the frequency of the measuring current.

R is the loop circuit resistance.

I_R is the current at resonance.

I_1 is any other current.

C_2 and C_1 are the values of capacitance of the loop circuit for which the current is I_1. (The inductance of the loop circuit is kept constant throughout.) Thus an increase in the capacitance variation required to reduce the current to a particular fraction of the resonant current (e.g., one-half) will cause an apparent increase in the loop circuit resistance.

In most cases the loop circuit resistance was constant within the accuracy of the method and the uniformity of conditions. Some variation was expected and experienced due to nearby

buildings, etc., increasing the losses in the loop circuit. Measurements of the resistance were made as a part of every field strength reading. Whenever the measurements indicated a resistance greater than a few per cent above normal, the reading was rejected since the energy input from the interference affected the accuracy of the reading. In almost all of these cases considerable interference could be observed by listening to the output of the set.

The procedure for a single measurement was as follows:

(1) The car in which the set was mounted was brought to the point chosen, and the motor stopped. The apparatus was set up,

Measurements of Field Strength:- Date:-5-17-26

Plate Voltage:- 45 Filament Current:- 0.25 Amp.

Weather:- Clear and Hot

Remarks:- Calibration Factor 0.88

		Park Trolley Bridge —		
Location	1	East Bank of River	Ardleigh St.	Sprague St.
Direction of Loop	2	80	150	175
Gain Step	3	1-4	1-5	1-4
Loop off	4	74.5	70	69
Meter Loop on	5	39.5	40	54
Diff:	6	35/53.5	30/56	15/65
Reading of Vernier 1/2 I	7	51	62	63
Condenser 1/2 I	8	27	36	28
Loop Resistance	9	11.7	12.6	17
Input-Calibration Curve	10	8.27	7.45	
Amplification Factor	11	6.5	2.6	
Actual Input Volts	12	53.7	19.4	
Field Strength	13	20,300 11.8	7,900 4.6	

Fig. 3—Sample Data Sheet

the filament current adjusted, and the normal reading of the detector-plate current meter noted to insure that the set and batteries were in proper condition.

(2) The detector meter was removed (since the jack connecting the meter simultaneously disconnected the last stage) and the station tuned in roughly by ear. The meter was then reinserted and the set was carefully tuned by adjusting the loop condenser, the oscillator condenser, and the direction of the plane of the loop. The degree of amplification was at the same time adjusted to bring the meter reading to a convenient portion of the calibration curve.

(3) The reading of the meter was taken with the loop connected and disconnected. (Lines 5 and 4 of Fig. 3.) The difference between these readings was observed, and from the calibration curve of the set (Fig. 2) the reading to indicate one-half of the

resonant current was obtained, and both these readings were noted (Fig. 3, Line 6). The calibrated micro-condenser in parallel with the loop condenser was then varied (on each side of the resonance value) until one-half of the resonant current was indicated, and the readings taken.

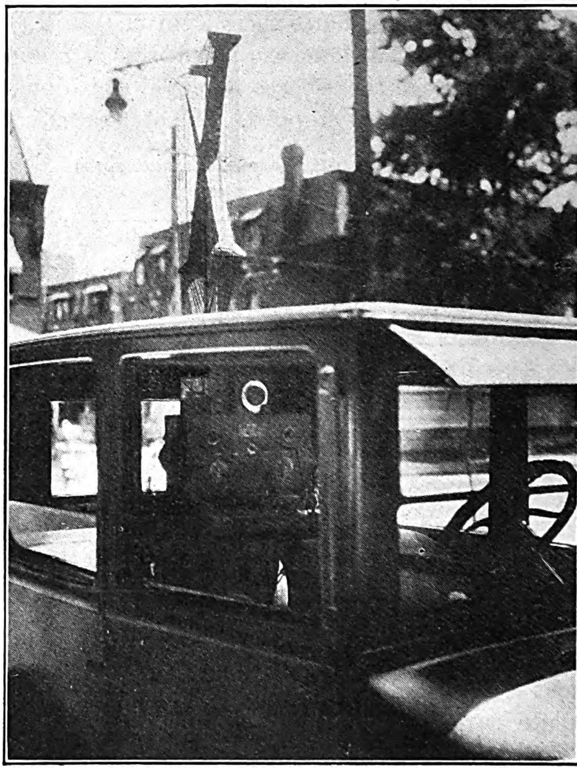

Fig. 4—Survey Car.

(4) The location, the direction of the loop with respect to the street and the amount of amplification were recorded. The delicate parts of the set such as the meters were then prepared for transportation.

A complete reading, depending on the skill of the operator, occupied from one to three minutes. Where interference was encountered, particularly of a transient nature, e.g., that due to trolley-cars, a longer period of time was required. Upon arrival

at the laboratory the loop resistance and the field strength were calculated.

Difficulties Encountered in the Field

The set was carried in a Ford sedan upon a double spring suspension above the rear seat (Figs. 4 and 5). The loop was mounted in a hole cut in the roof, with directional control and indicator within the car. During transportation the meters were disconnected, taken off suspension, and placed on the cushions of the rear seat, but despite the care taken, open circuits, broken tubes, and other mechanical injuries were frequent.

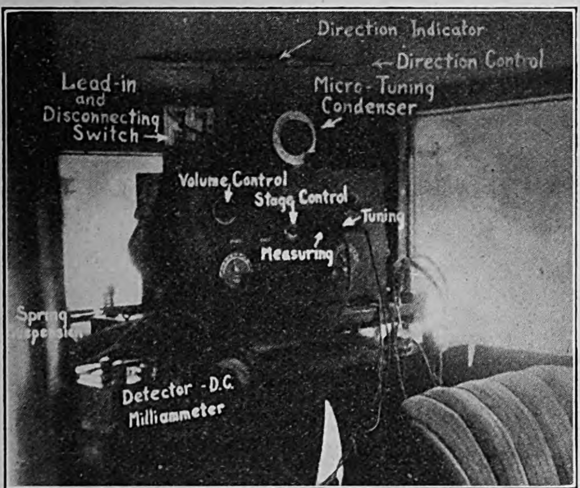

Fig. 5—Measuring Set Mounted in the Car.

Wherever possible measurements were taken in the center of open spaces or with some open space between the location and the nearest building in the direction of the transmitter. This was necessary since the shielding effect of even a two-story residence is considerable. On one occasion the transfer of the car to the opposite side of a seventy-foot street lined with two-story houses caused the field strength reading to double.

Field strength readings were spotted on a map of Philadelphia during the course of the field work. As the work progressed this afforded an opportunity of noting any unusual variations in an area and permitted the taking of check readings to determine the accuracy of such measurements. These local disturbances,

or shadows, were found to be due to such causes as the presence of large buildings, railways, elevated street car systems, or power lines. Since the object of the study was to get a general idea of

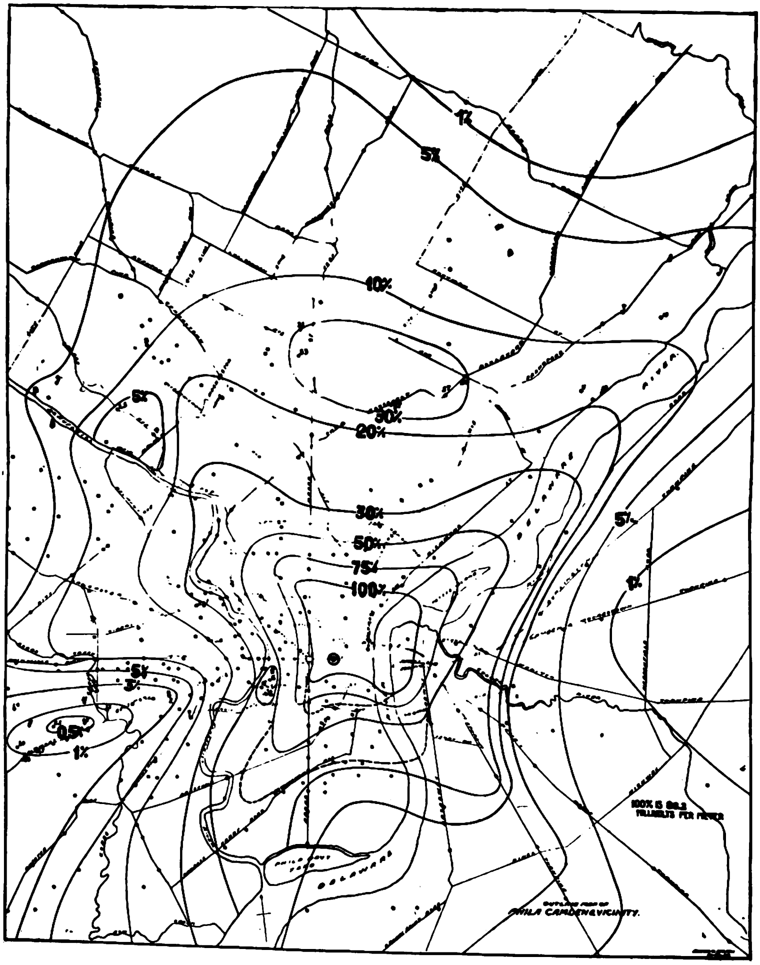

Fig. 6—Radio Field Strength Contour Map of Philadelphia

the distribution of radio waves over the city as a whole, and not to study conditions in a particular neighborhood, readings of this sort were discarded. In every case it was found possible to choose other locations in the immediate vicinity, which would not be subject to these local conditions.

It will be seen from the above discussion that, while a fair degree of accuracy was maintained, no great importance can be claimed for individual readings. Some interference was present at the majority of locations, there was always some effect from local buildings, and the readings taken on one side of the street would not necessarily hold for the other side. Readings taken at the street level would not be applicable to conditions on the roof of a three- or four-story building.

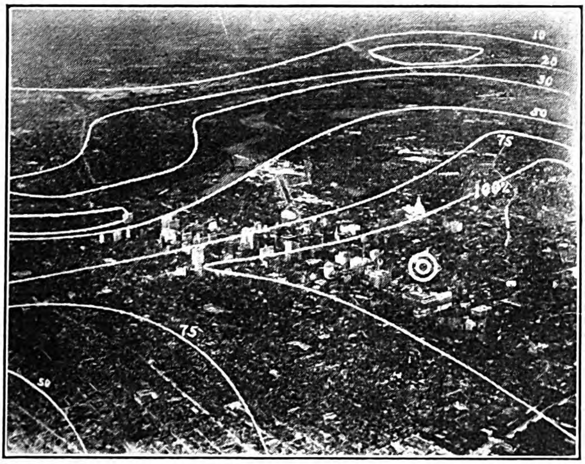

Fig. 7—Aerial Photograph of Philadelphia from Southeast of the Transmitting Station.

Discussion of Results

The most reliable of the readings taken were plotted on an outline map of Philadelphia (Fig. 6). The dots represent locations where measurements were made. Arbitrarily assuming 86.2 milli-volts per meter as 100 per cent, contour lines were drawn for various percentages. While it was impossible to draw these lines so as to be in accord with every measurement taken, a surprisingly large percentage of the measurements were consistent with the lines as drawn.

It will be seen from the map that the high percentage contour lines assume roughly the form of a square (instead of the oblong shape to be expected from an *L* type antenna) whose center is far to the northeast of the station. From Fig. 7 it will be seen that a line drawn through the station from northwest to southeast

passes to the north and east of practically all of the larger buildings of the city. It would thus be expected that the heaviest losses would occur to the south and west of the station, accounting for the shift of the center of area of the contour lines to the northeast. The very noticeable bulge in these lines toward the northwest can only be accounted for by the directional effect of the *L* type antenna. In the northeast and southeast directions from the station there are less pronounced bulges in the contours due to the Delaware River and to the heavily built section of Camden which is located directly east of the transmitter.

The solid contour lines all represent conditions at the time of completion of the survey. There was no change during the course of the survey except in the area south of Market Street and between the Delaware River and the Schuylkill. The check measurements made throughout the city shortly before the completion of the survey indicated that in this area the field strength had decreased considerably from the values found during the early months of measuring. The 100 per cent and 50 per cent contours existing at the start of the survey are plotted in dotted lines on the map. These contours have a rectangular shape, with the center almost due east of the station. During the course of the survey a large steel skeleton building was constructed directly south of and just across the street from the station, the roof being several stories above the antenna of WFI. The effect of this one large building close to the transmitting antenna is thus very marked.

The lower percentage contour lines show several interesting conditions. The most pronounced of these is a very dense shadow covering the whole of the area south of Market Street and west of the Schuylkill, the lowest field strength being found in the middle of Lansdowne, six miles west-southwest of the transmitter. Beyond this point there is a definite increase in field strength, shown to some extent by the contour lines drawn on the map, and even more markedly by measurements taken at locations beyond the confines of the map. This increase in field strength, or so-called "healing" of the shadow, is due to the fact that energy is fed in from other portions of the wave front. This effect checks the results of measurements taken in New York and elsewhere.

Shadows due to locally congested areas were found in two places. Below Market Street and skirting the east bank of the Schuylkill there is an area of decreased field strength due to the

presence of a group of factory and business buildings. Northwest of the station, between the 10 and 20 per cent lines, there is a larger area of decreased field strength due to the congested building area of Manayunk.

East of the Delaware River the shadow of the heavily built area of Camden is quite well marked. This shadow is nearly as dense but not so clearly defined as that in West Philadelphia.

Directly north of the station in the vicinity of the Roosevelt Boulevard between the 10 and 20 per cent lines there is an area of increased field strength. This would be expected from the fact

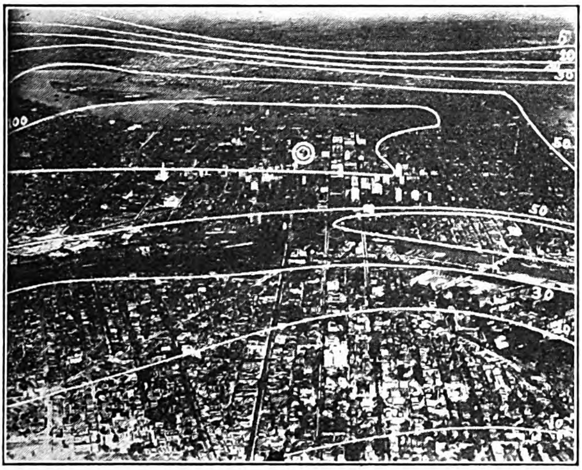

Fig. 8—Aerial Photograph of Philadelphia and Camden; West of Transmitting Station.

that in this vicinity there are only occasional groups of two-story houses, dotting a large area of open fields.

Whenever open country or rivers were encountered local bulges are found in the contour lines, indicating the decreased losses over such areas. This tendency of the lines is especially marked along the Delaware River northeast of the station; here the measurements taken as nearly on the river bank as possible indicate field strengths 50 to 100 per cent greater than those taken about a mile inshore. The percentages found at several of the locations in this area were noted on the map to illustrate this tendency.

All of the effects found above are consistent with existing theories of transmission. The pattern found for Philadelphia is

quite different from those of New York and Washington. While the New York pattern is characterized by its dumbbell shape and the Washington pattern by a circle, the Philadelphia pattern is roughly square, with its center far from the station. Other cities will undoubtedly exhibit equally interesting characteristic patterns.

Acknowledgment

The authors' thanks are due to the American Telephone and Telegraph Company not only for the loan of the equipment used in the measuring work, but also for advice and guidance received from its engineers. Thanks are also due to the Strawbridge and Clothier Company for cooperation in the actual measurements.

ON THE THEORY OF POWER AMPLIFICATION*

By

Manfred von Ardenne

(Berlin)

Summary—1. *In order to supply a maximum output wattage to an inductive loudspeaker by means of high-power tubes, regardless of the necessary amplitude of E_g, it is necessary to calculate the plate battery and grid-biasing potentials required to ensure distortionless reproduction.*

2. *From the inductive load on the plate circuit of the tube an elliptical shape of the working characteristic follows, referring to the $e_g; i_p$ diagram.*

3. *If the working ellipse is enclosed by a rectangular quadrangle, the sides of which are parallel to abscissa and ordinate axes respectively, then the slope or angle σ_{pz} of the diagonal of this quadrangle means the dynamic slope or mutual conductance of the tube for the given plate-circuit load, for*

$$\sigma_{p_z}=\frac{I_p}{E_g}$$

4. *By means of σ_{p_z} and starting from the conditions of distortionless reproduction the position of the diagonal of the quadrangle and its useful part are determined and furthermore the following equations for the minimum necessary plate-battery potential as well as for the necessary grid-biasing potential are found:*

$$\overline{E}_B=K+\overline{I}_p(r_p+R_b)+I_p\sqrt{(\omega L)^2+(r_p+R_{b_w})^2}=\overline{E}_g\cdot\mu\ .$$

$$\overline{E}_C=\overline{E}_G-I_p/\mu\sqrt{(\omega L)^2+(r_p+R_{b_w})^2}$$

5. *It is shown what conditions result if several loudspeakers are connected in series or in parallel with each other or if several tubes connected in parallel are employed to supply the loudspeaker.*

6. *Calculated and graphical solutions of the problem in question are given for practical examples and especially curves are shown illustrating the dependence of the effective output wattage of the loudspeaker on thei nternal plate resistance of the tube and on the operating frequency.*

I. Problem and Conditions

When it becomes desirable to obtain a maximum distortionless output of sound from a system of loudspeakers by means of a power-tube, then the question of calculating the necessary d-c. plate- and grid-potentials arises.

The condition of freedom from distortion comprises:

First, that in no case should the effective grid-potential of the tube, that is the d-c. grid-bias plus the impressed a-c. potential, reach the region of either appreciable grid-current or appreciable

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* Re-written in accordance with M. v. Ardenne, "Zur Theorie der Endverstärkung", *Jahrbuch*, Band 30, Heft 3.

curvature of the static tube characteristics; and, secondly, that the system of loudspeakers used should be capable of converting the impressed audio-frequency impulses into sound without distortion.

Furthermore, the condition is assumed to be fulfilled, as in practice is nearly always the case, that the preceding apparatus is capable of delivering any necessary amount of a-c. potential to the power-tube.

Roughly the load on the plate-circuit of the power-tube induced by the system of loudspeakers may be assumed to consist of an inductance L and an effective resistance R_{b_w}. This effective resistance R_{b_w} consists of the pure ohmic resistance R_b of the loudspeaker coils and the working resistance R_w, which itself again comprises both the losses resistance R_{w_v} and the useful acoustic radiation resistance R_{w_n}[1].

II. The Plate-Current

For the sake of simplicity it will be assumed that the power-tube is required to supply only a single loudspeaker.

The momentary plate-current then results from the steady direct plate-current and an a-c. value as indicated by:

$$i_p = \bar{I}_p + \tilde{i}_p \tag{1}$$

The steady direct plate-current may be taken to be:

$$\bar{I}_p = \sigma_{p_r} \cdot \left(\bar{E}_c + \frac{\bar{E}_B'}{\mu} \right) = \sigma_p \cdot \frac{r_p}{r_p + R_b} \cdot \left(\bar{E}_c + \frac{\bar{E}_B'}{\mu} \right) \tag{2}$$

In this formula σ_{p_r} designates the mutual conductance of the working characteristic, $\bar{E}_c$ the d-c. grid-biasing potential and $\bar{E}_B'$ the potential of the plate-potential battery, diminished by a correction potential K.

Furthermore

$$\tilde{i}_p = I_p \cdot \sin(\omega \cdot t - \psi) \tag{3}$$

can be written, if the impressed a-c. potential on the grid is sinusoidal, from which an also sinusoidal alternating plate-current follows. Since the tube may be looked upon as an a-c. generator

[1] R_{w_n} is in practice not absolutely constant, but changes according to the type of loudspeaker and is mainly dependent on the special resonance peaks of the latter. In the following R_{w_n} will be assumed to be constant, since for the usual loud-speaker-effectiveness its value may safely be disregarded.

with the e.m.f. or load-less potential $\mu \cdot Eg$, where Eg means the amplitude of the a-c. potential impressed on the grid,

$$I_p = \frac{\mu \cdot E_g}{\sqrt{(\omega \cdot L)^2 + (r_p + R_{bw})^2}} \tag{4}$$

follows; the denominator designates the impedance of the generator circuit. Finally the phase-angle between the plate-current and the grid-potential is given by

$$\psi = \text{arc tg}\left(\frac{\omega \cdot L}{r_p + R_{bw}}\right) \tag{5}$$

Thus the following final formula for the plate-current results:

$$i_p = \sigma_p \cdot \frac{r_p}{r_p + R_b} \cdot \left(\overline{E}_c + \frac{\overline{E}_B'}{\mu}\right) + \frac{\mu \cdot E_g}{\sqrt{(\omega \cdot L)^2 + (r_p + R_{bw})^2}} \cdot \sin\left[(\omega t) - \text{arc tg}\left(\frac{\omega \cdot L}{r_p + R_{bw}}\right)\right] \tag{6}$$

III. The Working Characteristic

By means of

$$lg = \overline{E}_c + E_g \cdot \sin(\omega t) \tag{7}$$

or

$$\sin(\omega t) = \frac{lg - \overline{E}_c}{E_g} \tag{8}$$

may be changed to define the relation between i_p and lg. For the sake of simplicity the letter α will be written instead of $\omega \cdot t$.

$$\sin(\alpha - \psi) = \sin\alpha \cdot \cos\psi - \sqrt{1 - \sin^2\alpha} \cdot \sin\psi \tag{9}$$

$$= \cos\psi \cdot \frac{lg - \overline{E}_c}{E_g} - \sin\psi \sqrt{1 - \left(\frac{lg - \overline{E}_c}{E_g}\right)^2} \tag{10}$$

$$= \frac{r_p + R_{bw}}{\sqrt{(\omega \cdot L)^2 + (r_p + R_{bw})^2}} \cdot \frac{lg - \overline{E}_c}{E_g} - \frac{\omega \cdot L}{\sqrt{(\omega \cdot L)^2 + (r_p + R_{bw})^2}} \times \frac{\sqrt{E_g^2 - (lg - E_c)^2}}{Eg} \tag{11}$$

From these equations the following results:

$$i_p = \sigma_{p_r} \cdot \left(\overline{E}_c + \frac{\overline{E}_B'}{\mu} \right) + \frac{\mu}{[(\omega L)^2 + (r_p + R_{b_w})^2]} \cdot [(r_p + R_{b_w}) \cdot (lg - \overline{E}_c)$$

$$-\omega L \cdot \sqrt{E_g^2 - (lg - \overline{E}_c)^2}] \qquad (12)$$

which may also be written thus:

$$y = A \cdot [a \cdot x - b \cdot \sqrt{E^2_g - x^2}] \qquad (13)$$

if the following substitutions are carried out:

$$\frac{u}{[(\omega \cdot L)^2 + (r_p + R_{b_w})^2]} = A; \quad \omega \cdot L = b;$$

$$r_p + R_{b_w} = a; \quad lg - \overline{E}_c = x; \quad i_p - \overline{I}_p = y. \qquad (14)$$

Since equation (13) can be brought into the form of the central equation of an ellipse

$$a_{1_1} \cdot x^2 + 2 \cdot a_{1_2} \cdot x \cdot y + a_{2_2} \cdot y^2 = k \qquad (15)$$

by means of the substitutions

$$a_{1_1} = A^2 \cdot (a^2 + b^2) = \frac{\mu^2}{[(\omega L)^2 + (r_p + R_{b_w})^2]};$$

$$a_{1_2} = -A \cdot a = \frac{-(r_p + R_{b_w}) \cdot \mu}{[(\omega L)^2 + (r_p + R_{b_w})^2]}; \quad a_{2_2} = 1; \qquad (16)$$

and

$$R = A^2 \cdot b^2 \cdot E_g^2 = \frac{\mu^2 \cdot E_g^2 \cdot (\omega L)^2}{[(\omega L)^2 + (r_p + R_{b_w})^2]^2}$$

it becomes apparent that the working characteristic of the tube for an inductive and ohmic load of the plate circuit is not a straight line, but, according to equation (12), an ellipse the axes of which form angles with the abscissa lg and the ordinate i_p.

IV. The Rectangular Quadrangle Enclosing the Working Characteristics

This formula leads, as will be shown in the following, to the equations desired expressing the necessary d-c. plate and grid potentials, if it is brought into relation with the conditions of the problem here under discussion which were stated above.

The working ellipse of the tube may be assumed to be enclosed by a rectangular quadrangle the sides of which run parallel to the ordinate and abscissa axes respectively and touch the ellipse. If, for loudspeaker reproduction, the regions both of appreciable grid-current and of excessive curvature of the static tube characteristics are to be avoided, then it is clear that neither the ellipse itself nor the quadrangle enclosing it may encroach upon these regions.

If the region of appreciable grid-current begins at the line $lg = \overline{E}_c$, which is at right angles to the abscissa lg, then the one

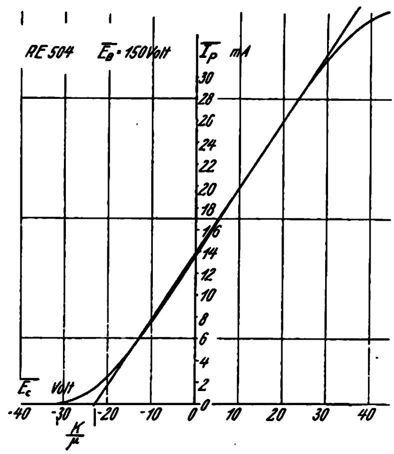

Fig. 1a

side of the quadrangle enclosing the ellipse must be identical with this line or, if considered necessary, correspond to even more negative grid-potentials.

The second side of the quadrangle running parallel to the first clearly lies at a distance of $2 \cdot Eg$ from the first. On the other hand, if the region in which the static characteristics are sufficiently straight reaches to $\overline{I}p_1$ and $\overline{I}p_2$ respectively, then the necessary position of the other two parallel sides of the quadrangle and their distance from each other is thereby defined.

V. The Diagonal of the Quadrangle

Since the respective lengths of the sides of the quadrangle accordingly are

$$lg_{max} - lg_{min} = 2 \cdot Eg \tag{17}$$

and

$$\overline{I_{p_1}} - \overline{I_{p_2}} = i_{p\max} - i_{p\min} = \frac{2 \cdot \mu \cdot E_g}{\sqrt{(\omega \cdot L)^2 + (r_p + R_{bw})^2}} \tag{18}$$

the diagonal of the quadrangle which passes through the point $\overline{E}_G; \overline{I}_{P_1}$ forms an angle with the abscissa lg, which may be expressed as follows:

$$\sigma_{p_s} = \frac{\mu}{\sqrt{(\omega L)^2 + (r_p + R_{bw})^2}} = \sigma_p \cdot \frac{r_p}{\sqrt{(\omega L)^2 + (r_p + R_{bw})^2}} = \frac{I_p}{E_g} \tag{19}$$

This angle here is expressed in the form of a mutual conductance, because, as will be shown later, the diagonal may with great advantage be assumed to be the working characteristic of the tube for an inductive and ohmic load on the plate-circuit, instead of the ellipse. The center of the ellipse and therefore also that of the enclosing quadrangle is the point $\overline{E}_c; I_p = \frac{\overline{I}_{p_1} + \overline{I}_{p_2}}{2}$; therefore this point also lies on the diagonal, which therefore possesses the equation

$$i_p - I_p = \sigma_{p_s} \cdot (lg - \overline{E}_c) \tag{20}$$

By means of equation (2) this may be changed to the following form:

$$i_p - \overline{I}_p = \sigma_{p_s} \cdot \left(lg - \frac{\overline{I_p}}{\sigma_{p_r}} + \frac{\overline{E_B'}}{\mu} \right) \tag{21}$$

VI. The Mathematical Solution of the Problem

Now the point with the abscissa

$$lg = \overline{E}_g$$

and the ordinate

$$i_p = i_{p\max} = \overline{I}_p + I_p$$

also belongs to the diagonal. If these values are substituted into equation (21), then after some operations the following formula results:

$$\overline{E}_B' = I_p \cdot (r_p + R_b) + I_p \cdot \sqrt{(\omega L)^2 + (r_p + R_{bw})^2} - \mu \cdot \overline{E}_G \tag{22}$$

Thus also:

$$\overline{E}_B = K + \overline{I}_p (r_p + R_b) + I_p \cdot \sqrt{(\omega L)^2 + (r_p + R_{bw})^2} - \mu \cdot \overline{E}_G \tag{22a}$$

If the values $\overline{E}_G$, $\bar{I}_p = \dfrac{\bar{I}_{p_1}+\bar{I}_{p_2}}{2}$ and I_p together with those of μ, r_p, and K are taken from the static tube characteristics and furthermore if the values $L_1 R_b$ and R_{bw} are known as the data of the loudspeaker in question, then the formula above supplies the desired value for the minimum amount of plate-battery potential necessary to meet the working conditions of the tube which were discussed above.

Furthermore, by means of

$$\overline{E}_c = \frac{\overline{I_p}}{\sigma_{p_r}} - \frac{\overline{E_B}'}{\mu} \tag{23}$$

this equation for $\overline{E}_B'$ leads to the following formula for the necessary d-c. grid-potential:

$$\overline{E}_c = \overline{E}_G - \frac{I_p}{\mu} \cdot \sqrt{(\omega L)^2 + (r_p + R_{bw})^2} \tag{25}$$

The a-c. potential impressed on the grid then must not exceed the value

$$\overline{E}_g = \frac{I_p}{\mu} \cdot \sqrt{(\omega L)^2 + (r_p + R_{bw})^2} \tag{26}$$

For ω a value of about $2\pi \cdot 800 = 5{,}000$ may be assumed, if the fact is taken into consideration that at the higher frequencies, with music as well as speech, the amplitudes of the impressed a-c. potentials are correspondingly smaller than at lower frequencies. The higher the frequency, the smaller the slope (mutual conductance) of the diagonal of the quadrangle which encloses the working ellipse becomes; and correspondingly smaller plate current variations are the result. If, as said above, the impressed a-c. potential amplitudes are greatest at a frequency of 800 Hertz, then in practice all those working curves for frequencies smaller than 800 must lie within the working ellipse for the frequency of 800 Hertz.

For the calculation of $\overline{E}_c$, however, a value corresponding to the smallest frequency in question must be assumed for ω, about $\omega = 2r \cdot 16 = 100$. Otherwise at these low frequencies the impressed a-c. potentials on the grid would reach the regions where the tube

characteristics are excessively bent. This will be shown by means of diagrams in section 13.

VII. Loudspeakers in Series and in Parallel

By connecting loudspeakers in parallel the impedance of the plate circuit is reduced, thus making smaller values of d-c. plate potential $\overline{E}_B$ and d-c. grid potential $\overline{E}_C$ sufficient; but at the same time the limit of a-c. potential which may be impressed on the grid is reduced.

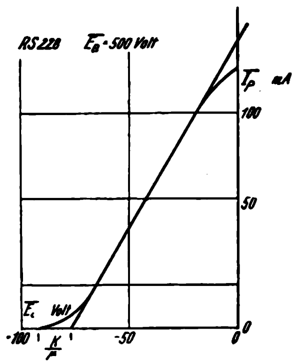

Fig. 1b

At the same time the acoustic wattage attainable is also reduced. This acoustic output wattage for a single loudspeaker corresponds to

$$W_\sigma = R_{w_p} \cdot \frac{I_p^2}{2} \tag{27}$$

while the attainable wattage with two loudspeakers of the same type in parallel is equal to:

$$W_\sigma' = 2 \cdot \left[R_{w_n} \cdot \frac{\left(\frac{I_p^2}{2}\right)}{2} \right] = \frac{1}{2} W_\sigma \tag{28}$$

If the loudspeakers are connected in series, then L and R_{b_v} become larger. Thus under these conditions both the d-c. plate and grid potentials have to be increased and also the amplitude of the impressed a-c. potential has to be correspondingly larger. The effective acoustic output wattage then, for the case of two similar loudspeakers connected in series, is equal to

$$W_\sigma'' = (2 \cdot R_{w_n}) \cdot \frac{I_p^2}{2} = 2 \cdot W_\sigma \tag{29}$$

Thus the output wattage is doubled, but of course only if, as said above, the d-c. plate and grid potentials as well as the impressed a-c. grid potential are correspondingly increased.

VIII. Practical Examples

A loudspeaker of the following electrical dimensions will be assumed:

$L = 2.35$ henrys; $R_b = 2000$ ohms; $R_{b_w} = 3000$ ohms.

In order to supply this loudspeaker with a maximum output wattage the tubes given in the table below will be considered; the values $\bar{I}_p$, I_p and $\bar{E}_G$ given in the table were taken from the static tube characteristics according to Figs. 1a–b.

TABLE I

Tube	E_f in volts	I_f in amps.	μ	r_p in Ohms	$\bar{I}_p$ in milli-amps.	I_p in milli-amps.	$\bar{E}_g$ in volts	K in volts
Telef. RE 504	3.5	0.5	4.5	7500	17	11	0	38.6
Telef. RS 228	7	1.1	5.5	3200	60	40	0	83.5

From these values follows:

TABLE II

Tube	$r_p + R_b$ in Ohms	Z_{b_w} in Ohms	Z_B in Ohms	$\bar{E}_b$ in volts	$\bar{E}_c$ in volts	E_g in volts
Telef. RE 504	9500	12130	15760	374	−38.2	25.4
Telef. RS 228	5200	12130	13300	927.5	−95.8	44.7
$\omega =$		5000	5000	5000	5000	100

In this table the different signs designate the data of the arrangement as follows:

$r_p + R_b$ = Internal plate resistance of the tube + pure ohmic resistance of the loudspeaker.

Z_{b_w} = Impedance of the loudspeaker system, $= \sqrt{(\omega \cdot L)^2 + (R_{b_w})^2}$

Z_B = Total impedance of the plate circuit, including the tube $= \sqrt{(\omega \cdot L)^2 + (r_p + R_{b_w})^2}$

$\bar{E}_B$ = Minimum necessary plate-battery potential, chosen with regard to the higher frequencies, $\omega = 5{,}000$.

$\bar{E}_C$ = Minimum necessary negative grid-biasing potential, chosen with regard to the higher frequencies, $\omega = 5{,}000$.

E_g = Maximum value of impressed a-c. grid-potential permissible, defined with regard to the lower frequencies, $\omega = 100$.

IX. The Use of Several Power Tubes Connected in Parallel

According to the above calculations a clearly defined minimum plate-battery potential is necessary to obtain a maximum output wattage from a given power-tube to a given loudspeaker. This fact might lead to the false assumption that, in order to obtain an equal output wattage it would be possible to substitute a number of smaller tubes connected in parallel for a single but larger power-tube and at the same time using less plate-battery potential for the several smaller tubes.

If it is assumed that n similar, smaller tubes are connected in parallel, each one of which possesses the data $\overline{E}_f'$, $\overline{I}_f'$, μ', r_p', $\overline{I}_p'$, I_p', K', then these tubes correspond to a single hypothetical tube of the following electrical dimensions:

$$\overline{E}_f=\overline{E}_f',\ \overline{I}_f=n\cdot\overline{I}_f',\ \mu=\mu',\ r_p=\frac{r_p'}{n},\ \overline{I}_p=n\cdot\overline{I}_p',\ I_p=n\cdot I_p',\ K=K'$$

In this case the different electrode-systems may be assumed to be enclosed in a single evacuated glass tube.

The necessary plate-battery potential for the n power tubes connected in parallel then is

$$\overline{E}_B=K'+n\cdot\overline{I}_p'\cdot\left(\frac{r_p'}{n}+R_b\right)+n\cdot I_p'\sqrt{(\omega L)^2+\left(\frac{r_p'}{n}+R_{b_w}\right)^2} \quad (30)$$

For example, with four tubes RE 504 arranged in this manner and with the following electrical dimensions of the circuit

$L=2.35$ henrys; $R_b=2{,}000$ ohms; $R_{b_w}=3{,}000$ ohms

the necessary value of plate-battery potential resulting would be 861 volts. The output current of these four tubes is $\overline{I}_p=4.17=68\ m_A$ and $I_p=4.11=44\ m_A$ for a filament wattage consumption of $4\cdot3.5\cdot0.5=7$ watts. About the same output current is supplied by a single tube of the type RS 228, namely $\overline{I}_p=60\ m_A$ and $I_p=40\ m_A$, for a filament consumption of approximately the same value, namely $7\cdot1.1=7.7$ watts. This tube requires a minimum plate-battery potential of 927.5 volts according to Table II.

X. The Dependence of the Plate Current and Useful Output Wattage from the Internal Plate Resistance of the Tube and from the Operating Frequency

If the load on the plate circuit, that is the values L, R_b, and R_{b_w} of the loudspeaker system, is assumed to be constant and predetermined, then the plate-current is, if a certain amplitude

E_g of impressed a-c. grid-potential is predetermined, only dependent from the amplification factor μ of the tube, which may be regarded as a constant, from the internal plate resistance of the tube and from the operating circuit frequency ω. According to equation (19), in which σ_{p_z} is a dynamic mutual conductance,

$$I_p = E_g \cdot \sigma_{p_z} = \frac{\mu \cdot E_g}{Z_B} = \frac{\mu \cdot E_g}{\sqrt{(\omega L)^2 + (r_p + R_{b_w})^2}}$$
$$= E_g \cdot \sigma_p \cdot \frac{r_p}{r_p + R_{b_w}} \cdot \cos\psi \tag{31}$$

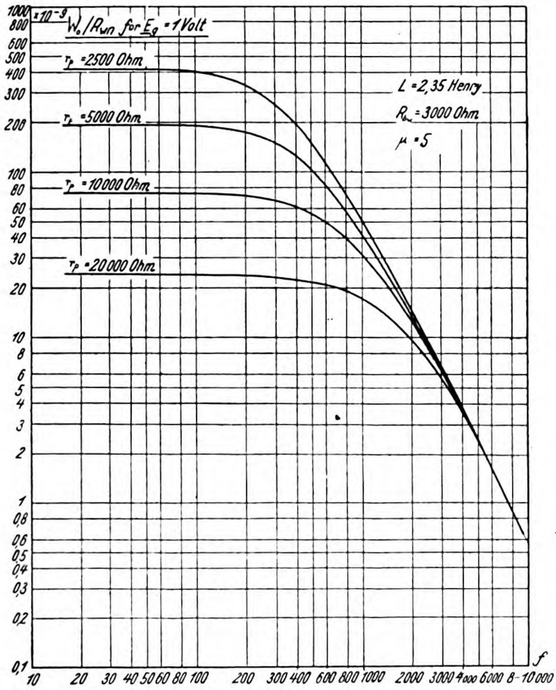

Fig. 2

By this means follows:

$$W_\sigma = \frac{1}{2} \cdot R_{w_p} \cdot I_p^2 = \frac{1}{2} \cdot R_{w_p} \cdot E_g^2 \cdot \left[\sigma_p^2 \cdot \left(\frac{r_p}{r_p + R_{b_w}} \right)^2 \cdot \cos^2\psi \right] \tag{32}$$
$$= \frac{1}{2} R_{w_p} \cdot E_g^2 \cdot \frac{\mu^2}{(\omega L)^2 + (r_p + R_{b_w})^2}$$

This dependence of the output wattage supplied to the loudspeaker on the internal resistance of the tube and on the frequency

is shown for the values given in the above equations by a number of curves in Fig. 2.

In judging these curves it must be borne in mind that they only represent the electrical output wattage supplied to the loudspeaker, which the latter converts into sound, but not the actual acoustic air pressure amplitudes which are produced by the mechanical vibrations of the loudspeaker diaphragm at different frequencies. The actual radiated sound amplitudes, however, increase with the frequency in accordance with an approximately square law for equal amplitudes of mechanical vibration. This (if the losses are disregarded) practically square law approximately corresponds to the opposite incline of the curves shown above at higher frequencies and thus practically equal volume of sound is obtained at the higher frequencies. At the lower frequencies, however, matters become much more complicated and will therefore not be discussed here.

XI. A Contribution to the Designing of Tubes for Power-Amplification Purposes

In order to design suitably a tube meant for power-amplification one may start with the condition that for a certain plate-battery potential $\overline{E}_B$ a certain a-c. amplitude I_p is to be supplied to the loudspeaker. According to equation (22)

$$\overline{E}_B - K = I_p \cdot (r_p + R_b) + I_p \cdot \sqrt{(\omega L)^2 + (r_p + R_{b_w})^2}$$

In this equation K is always defined with sufficient exactness for preliminary calculations (refer to Table I) as

$$K = 0.1 \cdot \overline{E}_b.$$

Furthermore

$$\overline{I}_p = M \cdot I_p.$$

may be written and for M the value 1.5 may be substituted in the case of high-power tubes and 2 in the case of tubes of ordinary dimensions.

By this means

$$\frac{\overline{E}_B}{I_p} = \frac{M}{0.9} \cdot (r_p + R_b) + \frac{1}{0.9} \cdot \sqrt{(\omega L)^2 + (r_p + R_{b_w})^2} \tag{33}$$

is found. Now if, as mentioned above, $\dfrac{E_B}{I_p}$ is predetermined, if a

suitable value is substituted for M (between 1.5 and 2) and if average values for L, R_b and R_{b_w} corresponding to a good loud-speaker available, are substituted, then the above equation will supply the desired optimum value for the internal plate resistance r_p of the tube. This (maximum) value of r_p the tube must have, if the desired output wattage defined by I_p is to be obtained at the plate battery potential $\overline{E}_B$ without the regions of appreciable grid-current or excessive curvature of the static tube characteristics being entered into. As is apparent, r_p *must be the smaller,*

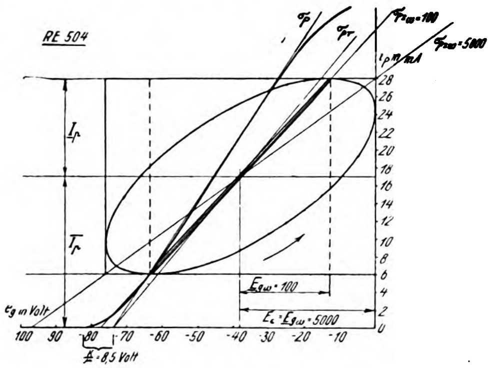

Fig. 3

the smaller the plate-battery potential and the greater the output wattage desired are.

If in equation (33) $\frac{\mu}{\sigma_p}$ is written instead of r_p as follows;

$$\frac{\overline{E}_B}{I_p}=\frac{M}{0.9}\cdot\left(\frac{\mu}{\sigma_p}+R_b\right)+\frac{1}{0.9}\sqrt{(\omega L)^2+\left(\frac{\mu}{\sigma_p}+R_{b_w}\right)^2} \qquad (34)$$

then this equation is the means of showing how the amplification factor μ of a certain tube must be changed, at the same time retaining the original mutual conductance, if the tube does not fulfil the conditions imposed regarding $\overline{E}_B$ and I_p; which latter is for example the case, if at the maximum plate-battery potential allowable a portion of the straight part of the static tube characteristic still runs into the region of positive grid-potentials. A small value for $\frac{\overline{E}_B}{I_p}$, that is a high specific efficiency of the tube, is

obtained with low values of the amplification factor μ of the tube, which may be as small as 2.5.

The fact, however, must not be overlooked that the reduction of μ reduces the amplification of the tube and necessitates greater amplitudes of impressed a-c. grid-potential in order to retain the original output wattage required. For

$$W_\sigma = \frac{1}{2} \cdot R_{w_p} \cdot (\mu \cdot E_g)^2 \cdot \frac{1}{(\omega L)^2 + (r_p + R_{b_w})^2} \tag{35}$$

Thus E_g must be changed in porportion with $\frac{1}{\mu}$, that means that the e.m.f. must be constant if the output wattage is to remain unchanged.

XII. Explanation of the Problem by Means of Diagrams

In Figs. 3 and 4 the practical cases corresponding to Tables I and II are diagrammatically reproduced. It is probably sufficient to explain one of these diagrams, because they are in principle exactly similar and only differ in certain values.

In Fig. 3, which refers to the tube RE 504, at first the diagonal of the rectangular quadrangle enclosing the working ellipse was drawn through the point $\overline{E}_c - E_{g_{\omega=5000}} = 0$; $\overline{I}_p + I_p = 28\text{m}_\text{A}$ according to a calculation of $\sigma_{p_{s\omega=5000}}$ from equation (19). The crossing-point of this diagonal with the straight line $\overline{I}_p = 17\text{m}_\text{A}$ results in the finding of the length $\overline{E}_c = E_{g_{\omega=5000}}$, which represents the necessary d-c. grid-biasing potential of 38.2 volts negative. Following this the rectangular quadrangle itself was drawn by means of $\overline{I}_p - I_p = 6\text{m}_\text{A}$ and the known opposite corner of the quadrangle. Thus also, by means of equations (13) and (14) the working ellipse for the frequency of $\omega = 5000$ enclosed by the quadrangle could be drawn. The working direction of the ellipse is indicated by an arrow, for a capacitative load it would be in the opposite sense.

Thereupon a second diagonal was drawn with the angle or slope (mutual conductance) $\sigma_{p_{s\omega=100}}$; by means of this diagonal the second and smaller quadrangle drawn in broken lines was defined, which determines the value $E_{g_{\omega=100}} = 25.4$ volts and which encloses a shorter and much narrower ellipse corresponding to a frequency of $\omega = 100$. It is apparent from this narrow ellipse that if a-c. potentials of a frequency of $\omega = 100$ were impressed on the

grid exceeding the value $E_{g\omega=100}$, then the narrow ellipse would be lengthened at both ends and would thus enter into the regions of excessively bent static tube characteristics determined by the plate-current limit values $\bar{I}_p+I_p$ and $\bar{I}_p-I_p$. This of course must be avoided if the fundamental conditions imposed are to be fulfilled. It is due to this fact of the tube being modulated up to the limit $\bar{I}_p+I_p$ by small amplitudes of lower frequency in the same degree as by large amplitudes of higher frequency that there is an unequal frequency response of the plate-current and the output wattage, as shown by the curves reproduced.

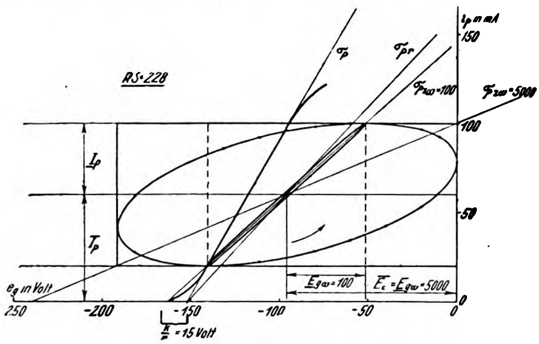

Fig. 4

Finally a third straight line is drawn through the point $\bar{E}_c;\bar{I}_p$ this time with the angle σ_{p_r} and through the crossing-point of this line with the abscissa a fourth straight line is drawn with the angle σ_p. In accordance with this last straight line a static tube characteristic is also drawn into the diagram. From this latter curve $\dfrac{K}{\mu}=8.5$ is found to be. If furthermore the tube characteristic drawn for a plate battery potential of 150 volts shown in Fig. 1 is assumed to be included in Fig. 2, then the distance between the characteristic already present in this diagram from the characteristic just transferred, measured parallel to the abscissa in volts, must be $\dfrac{\bar{E}_B-150}{\mu}$. From this value a minimum necessary plate battery potential of $\bar{E}_B=374$ volts follows.

Author's Note: The nomenclature employed in this paper corresponds to that given by E. Leon Chaffee in his paper "Vacuum Tube Nomenclature."

CONDENSER SHUNT FOR MEASUREMENT OF HIGH-FREQUENCY CURRENTS OF LARGE MAGNITUDE

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Summary—*The necessity for an accurate ammeter for large high-frequency currents is pointed out. A new device consisting of a large condenser in parallel with a small condenser, and the latter carrying the current to a small thermocouple ammeter, is described.*

A device of this nature can be made very accurate; in fact, comparable in accuracy to any available standards.

The construction of the device includes provisions for reducing and restricting the electrostatic and electromagnetic field, due to large current, the reduction of distributed inductance and capacity, and a provision to prevent the resonance effect of high harmonics of the operating current. Provisions are also made for locating the measuring instrument at a distance from the circuit. Large ratings are possible by connecting a number of condenser units in parallel.

THE use of large broadcasting stations and other continuous-wave, high-power installations has created a demand for accurate means of measuring high-frequency current of large magnitude.

The methods so far in use are all limited in one particular or another. The use of the hot-wire expansion type instruments is not feasible for values above 10 amperes, as the size of heating element becomes excessively large and the skin effect does not allow the subdivision of hot wire into parallel elements.

The direct thermocouple type has been used with satisfactory results up to currents of 100 amperes, but the heating element of the higher ranges becomes bulky and expensive to build on account of the large-sized conductors and careful workmanship required. Also the skin effect becomes appreciable at the higher frequencies.

An iron core transformer for reducing the high-frequency current so that it can be applied directly to a small instrument gives satisfactory results for frequencies up to 500 kc. For higher frequencies, the heating of the iron parts of the transformer becomes quite appreciable and is the greatest drawback. At 2,000 kc. and above, it is difficult to use such a transformer; the heating of parts, the influence of stray fields, and the distributed capacity of windings become quite objectionable.

* Original Manuscript Received by the Institute, August 10, 1927.

This article describes a novel arrangement which will permit the limits of operation to be extended as far as the present art of radio transmission requires.

The advantage of the new condenser type of ammeter for large currents lies in its accuracy and simplicity, combined with

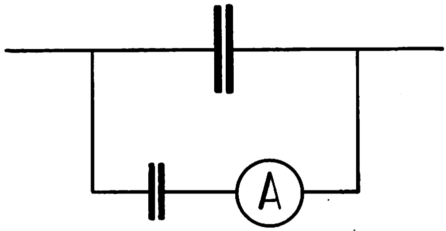

Fig. 1

its comparatively low cost, even for the highest of frequencies.

Fig. 1 illustrates the method by which currents of large magnitude and high frequency can be satisfactorily measured. It consists in general of two condensers in parallel; a large one

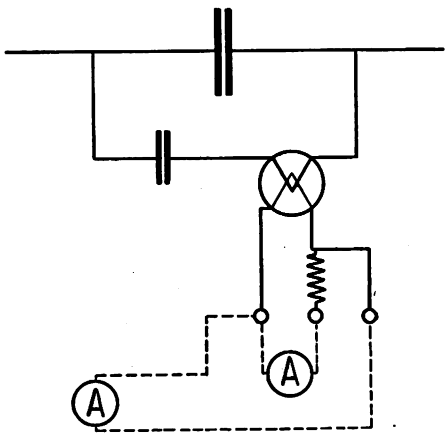

Fig. 2

which carries the greater portion of the current to be measured; without appreciable voltage drop, and so constructed that it can pass large current at high frequency without appreciable losses, and one considerably smaller, designed to shunt off a predetermined fraction of the total current through a small ammeter, either of the hot-wire or the thermocouple type. In the latter case, the meter may be located at a distance from the main circuit, as illustrated in Fig. 2.

A device of this nature, if properly designed, will give satisfactory measurements of current at frequencies as high as 60,000 kc., which is practically the limit of the present-day operations of radio stations. It can be designed for higher frequencies. The error due to the resistance of the thermocouple element is practically negligible, even at the highest frequencies used. Thus, if the condenser has a capacity of 0.001 μfd. and the thermocouple element has a resistance of 2.6 ohms, the error becomes one-half of one per cent at 6,000 kc. (50-meter wavelength). For shorter waves, smaller capacity would be used.

The only source of error actually found in operation is due to the fact that the two condensers and the thermocouple form a

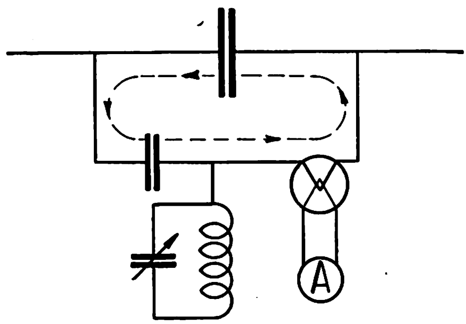

Fig. 3

closed circuit which has a resonant frequency that sometimes comes within the range of some harmonic of the frequency of operation of the instrument. The presence of such a condition becomes apparent in an obvious irregularity of the meter reading. Fig. 3 shows a very evident method to avoid this error due to the resonant frequency. An auxiliary circuit which is tuned to the resonant frequency of the closed circuit referred to is connected to some point of the condenser shunt and actually absorbs the power of the harmonic from this circuit and in this way eliminates this error. Since this tuned circuit is connected only at one point, its effect at all other frequencies is entirely negligible.

It is worthy of notice that the accuracy of this instrument cannot really be checked by any available standards of high-frequency current. Probably the most accurate fundamental method of measuring high-frequency current is by means of the calorimeter ammeter in which the heating due to the current registers the value of that current in terms of the resistance of

the heating element. Even this method, however, is subject to two errors which are difficult to eliminate. One is the actual value of resistance at the high frequency and the second is the distributed capacity of the heating elements and the calorimeter apparatus.

If it is remembered that the capacity values used in condenser shunt are considerably in excess of any distributed capacity, and moreover, with a properly constructed mica condenser, these values are constant at all frequencies, and if it is further realized

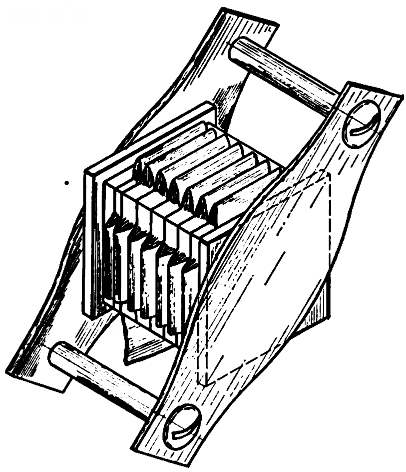

Fig. 4

that the distributed inductance and resistance of leads are really negligible, the accuracy of this method becomes self-evident; it establishes a standard of large high-frequency current measurement determined only by the accuracy of the meter element in series with the small condenser.

Fig. 4 illustrates an early design of condenser element which was found suitable for this apparatus. It is a unitary structure with a powerful clamp and two capacity elements, both within this clamp. One element consists of a number of metal foils in parallel, and gives the large capacity, while one extra foil brought out as a separate lead gives a small capacity. It is evident that the construction is made so symmetrical that there is no chance of one capacity changing relatively to the other. It will also be seen that the incoming and outgoing leads of this condenser are on the same side of the clamp. There is therefore no magnetic loop

around the path of the current, and consequently an important cause of the losses is eliminated.

Fig. 5 is an illustration of a meter of this type constructed for operating with a current of 100 amperes. It will be seen that there

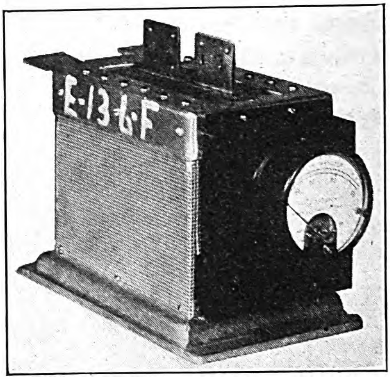

Fig. 5—External Appearance of an Early Model of Condenser Shunt.

are two leads coming out through the cover which can be connected in parallel or individually, depending on the current to be measured. There are two condenser elements corresponding to these leads. Only one of these condenser elements contains a small

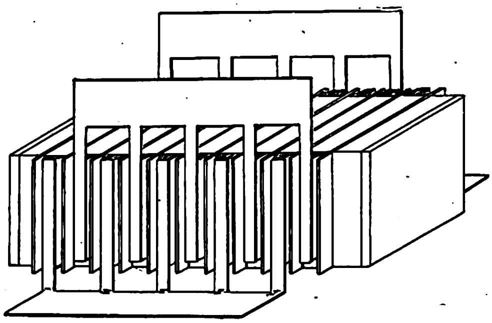

Fig. 6

capacity in series with the meter, that is, the element closest to the meter end. If this element alone is used, the reading is 50 amperes. If the second element is also connected, the reading becomes 100 amperes, and the meter readings must be multiplied by 2 without affecting the calibration of the instrument.

It is a well-known fact that a large current flowing at very high frequency causes considerable losses, and the heavier the current the more the relative losses. The cause of these losses was thoroughly analyzed and it was found that the probable reason for them is in the electro-magnetic field surrounding any path carrying large values of current. Such a magnetic field at very high frequency undoubtedly sets up an electro-static field through the body of the insulation and this electro-static field in turn causes dielectric losses. The problem therefore reduces to elimination of the magnetic field so as to eliminate the consequent electro-static field.

Fig. 6 shows diagrammatically a construction of a condenser where this electro-magnetic effect is practically entirely eliminated.

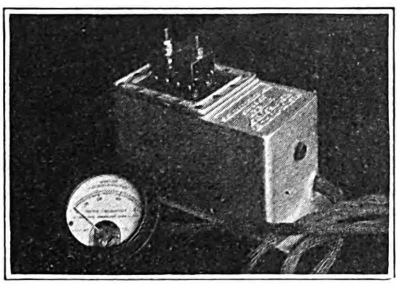

Fig. 7—Recent Model of Condenser Shunt with Shielded Lead and Panel Type Instrument

It will be seen that the condenser consists of a number of rather narrow sections interleaved with each other and so connected that the current through the body of the condenser has two opposite paths, each carrying an equal amount of current. Thus, the magnetic effects of the two paths cancel each other and the resulting electro-magnetic field is confined to the immediate vicinity of the conductors in each individual section. It was found by actual experiment that the losses by this construction were tremendously decreased so that a condenser which was previously giving a temperature rise of 15 or 20 deg. C. would with the new construction give an inappreciable temperature rise of 2 or 3 deg.

It will be further seen that this type of construction permits the use of a simple clamp surrounding the condenser, with leads coming out on both sides. Since the current now forms two loops in opposite directions, there will be no magnetizing effect on the

clamp. If, further, bronze springs and brass clamping rods are used, the losses in the clamp are practically eliminated. As before, the small capacity is introduced by an extra foil in the condenser. It was found desirable to split this condenser into two parallel sections so that when a smaller current rating is required from the condenser shunt, half of the capacity may be used, giving full scale rating on the meter at one-half the current.

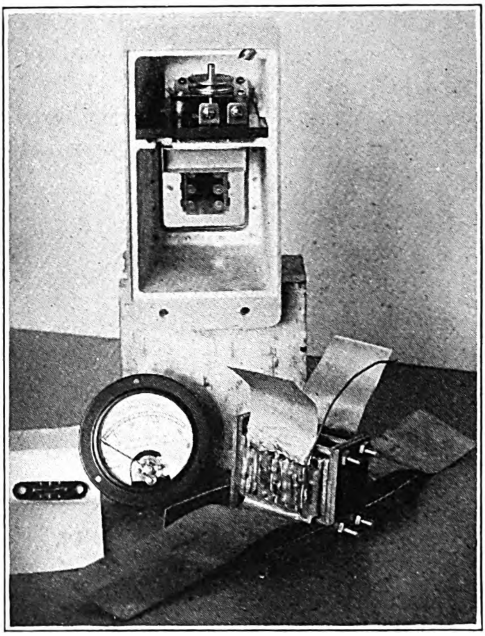

Fig. 8—Constructional Details of Condenser Shunt.

A typical example for a condenser to operate at 50 amperes is as follows. The shunt element is 0.199 μfds. and the part constituting the small condenser is 0.001 μfd. At 6,000 kc., the potential across this condenser will be only 6.6 volts. With this condenser a ratio of current is obtained of 200-to-1, so that with a 50-ampere condenser shunt the thermocouple will carry one-quarter of an ampere. If one-half of this condenser is used, as described above, to get the 25-ampere rating, the ratio is 100-to-1, giving again a current of one-quarter ampere for 25 ampere through the condenser shunt.

Fig. 7 illustrates a condenser of this type as it appears from the outside. In this case it was found more advantageous for commercial reasons to mount the instrument itself at a distance from

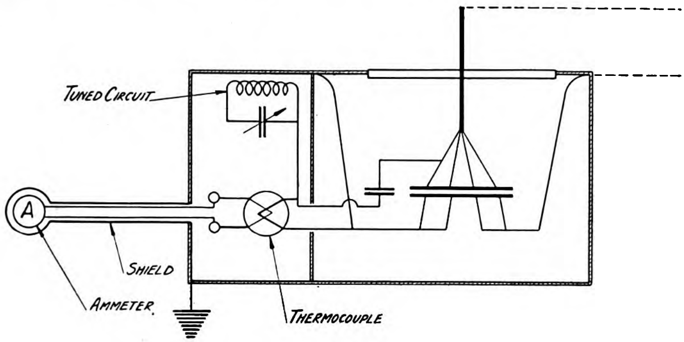

Fig. 9

the condenser shunt, and provide a shielded lead between the instrument and the condenser.

A split terminal is brought out on the top with a paralleling strap between the two halves. For currents more than 50 amperes,

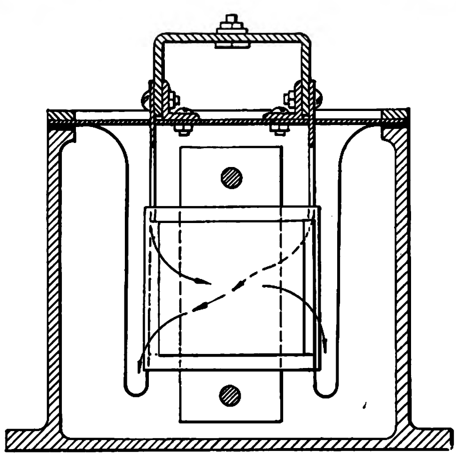

Fig. 10

a number of such condenser units may be paralleled, but of course only one instrument will be required. In that case it is evident that the ratio would be multiplied by the number of condenser units used.

Fig. 8 shows the inside appearance of this structure with a condenser element as described above in one compartment, and the small resonant circuit with the thermocouple in another compartment. A hole for bringing in a screw driver and adjusting the resonant circuit is sealed after the instrument is assembled, while a new thermocouple may be replaced by removing the lower cover.

Fig. 9 shows a diagram of connections inside of the condenser shunt. It will be seen that the lead from the small condenser to

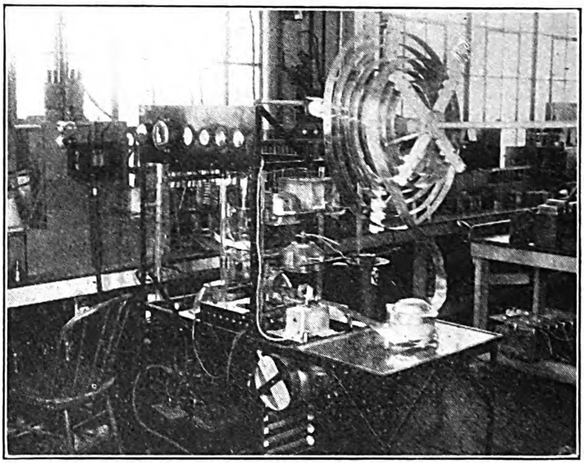

Fig. 11—Short Wave Radio Testing Set with Condenser Shunt Installed

the thermocouple is made as short as possible so as to avoid the inductive effect of this lead in the closed circuit consisting of two condensers and the thermocouple. This reduces one inductive effect and thereby increases the resonant frequency of that circuit until it is effective only at a very high harmonic of the operating current. This is then compensated by the tank effect of the compensating resonant circuit.

Fig. 10 shows the actual location of the main leads in the condenser element. Attention is drawn to the fact that by interlacing the condenser sections and by arranging ingoing and outgoing connections through two parallel leads the magnetic effect of the current outside of the condenser element is almost entirely eliminated and to further avoid this magnetic effect on the thermocouple the condenser element is enclosed in a metal partitioned compartment.

Fig. 11 illustrates the location of this instrument on a radio transmitting set. The condenser shunt is mounted in a convenient location next to the ground lead, while the instrument is located on the instrument panel, with a shielded lead between the instrument and the condenser. In this particular set, which was designed for the purpose of testing condensers, and is capable of operating on a wide range of frequencies from 1000 meters to 20 meters, the condenser shunt was found to give correct readings for the full range and for currents in excess of 50 amperes an additional parallel unit could be applied.

A meter similar to the one described above has been in continual use for over two years on a testing set where the frequencies have ranged from 100 kc. to 6,000 kc., and the current values have ranged up to 120 amperes. It has been found that the meter indications are consistent and reliable at all these values. In fact, its accuracy has been such that it was possible to measure the voltage in a circuit by connecting such an ammeter in series with a known condenser, and determining the voltage drop in the condenser by calculation.

For developing a special thermocouple ammeter used in connection with the condenser shunt, an acknowledgment is due the Weston Electrical Instrument Corporation of Newark, New Jersey.

BOOK REVIEWS

Dielectric Phenomena: Electrical Discharge in Gases, BY S. WHITEHEAD. D. VAN NOSTRAND COMPANY, INC., New York, N. Y. 176 pages. Price $4.00.

The scope of this book is outlined in the preface by E. B. Wedmore. It is a scientific summary of what is known of electrical discharge in gases, as the sub-title indicates. The main title is to be justified by further publications dealing with liquids and solids.

The volume is written in three parts. The Introduction is a brief outline of the theory of ionic conduction through gases, such topics as "Mechanics of the Atom and Molecule," "Ionisation by Impact," "Radiations Emitted by the Discharge," and "Townsend's Continuity Theorem" being epitomized. In the second section the physics of electrical sparking is given, while the third portion of the book is concerned with corona phenomena. The term corona is taken as including dark, glow, and brush discharges. Mr. Whitehead's work is limited to "electrical discharges through gases at pressures for which the mean free path is small compared with the electrode dimensions," thus excluding discharges in vacua, and the electric arc. An arc is understood as an electrical breakdown between conductors in a gas, the electrodes themselves becoming sufficiently hot to contribute to the phenomena following the initial discharge.

The discussion is in the field of physics rather than engineering. The mathematical treatment of the various topics is adequate but not prohibitively difficult. For those readers who desire graphical results there are forty-two diagrams showing the variation of gradients and other effects. An engineer who picked up the book with the object of reading about horn gaps, the proportioning of the V-shaped sides, and the amount of energy which they may be depended on to break, would be disappointed in his search for specific engineering data, but he could learn how much the spark voltage of the sphere gap is reduced by rain or other water on the surfaces, and that humidity appears to have little effect, whereas the needle gap is markedly influenced by humidity. At the end of the book he would also find an alphabetically arranged bibliography covering the subject from 1860 to very recent publications. There is also a rather brief index.

Original Manuscript Received by the Institute, November 28, 1927.

While Mr. Whitehead's résumé is hardly of interest to radio engineers generally, it should be read by transmitter design specialists, who will find it to their advantage to substitute for isolated empiricisms a logical physical treatment of high voltage discharges.

CARL DREHER

The Propagation of Radio Waves Along the Surface of the Earth and in the Atmosphere, BY PROFESSOR P. O. PEDERSEN. 244 pages and appendix. Published by "Danmarks naturvidenskabelige Samfund" and sold by G. E. C. Gad, Vimmelskaftet 32, Copenhagen K. Denmark. Price 15 Kr. (about $4.00).

In this book the author has attempted to give a connected physical theory of radio wave propagation—and it is no doubt the first case where really all the well-known facts of radio propagation have been assembled and brought into connection with the corresponding fields of physics.

It is first pointed out that only an ionization theory can satisfactorily explain the conditions of radio transmission, which is so extremely variable with the daytime, season, and wavelength. In the following chapters the physical basis of such a theory is worked out, and it is shown that radio experience throws light on many problems concerning the atmosphere, the laws of ionization and recombination, etc.

In chapter IV the pressure and composition of the atmosphere are treated. It is shown that no appreciable quantities of hydrogen can be present in the higher atmosphere, which therefore consists merely of helium—together with small percentages of nitrogen and oxygen; and that radio experience determines quite precisely the air pressure between 80 and 160 km. altitude. This determination would probably be in good agreement with the mass-density resulting from Lindemann and Dobson's investigations of meteors, if some errors in their theory are corrected.

Chapter V deals with the ionization of the atmosphere. The coefficient of recombination is shown to be proportional to the air pressure according to the theory of Langevin but for very low pressures will approach a constant value of about 10^{-5} times that at 760 mm Hg pressure, as shown in a theorem of the author. The different possible sources of ionization are now investigated. The ultra-violet radiation from the sun is the chief determining

Original Manuscript Received by the Institute, December 22, 1927.

factor though radiation from stars of very high temperature also has some influence on the propagation of very long waves. Formulas are developed for the numbers of free electrons and of mono-molecular and complex ions per cc. by day and by night.

In chapters VI and VII is determined the influence of electrons and ions on the dielectric constant and the conductivity of the atmosphere with or without the magnetic field of the earth acting, and taking into consideration the collisions between the different charged and uncharged particles. The reader will find several new and exact formulas set forth in this part of the book. In chapter VIII the dependency of the refractive index and the attenuation constant on the dielectric constant and the conductivity are treated. The approximate formula $n=\sqrt{\epsilon}$ is shown to be very unsatisfactory in many cases, and the exact formulas, taking into consideration the influence of both the dielectric constant and the conductivity, are given and represented in curves and charts.

For any set of values of the refractive index and the attenuation constant expressed as functions of the height over the earth and of the frequency we may determine the path and the attenuation of rays leaving the earth under a given earth angle according to the methods developed in chapter X. It is pointed out that the ray path will depend on both the earth angle and the frequency. This section of the book gives a very striking description of what happens to a ray penetrating into an ionized part of the atmosphere and indicates the average picture of radio propagation resulting from any ionization theory.—In other different theorems the influence of the refractive index and the attenuation constant on the rotation of the plane of polarization, on the phase- and group-velocity, etc., is treated.

In chapters IX and XI, the fundamental theories developed in the foregoing chapters are used for determining the actual state of ionization and for checking the theory against experience. Here are also used the old simplified theories of propagation listed in chapters I–III and the very useful formulas for reflection at the surface of the earth developed in chapter VIII. The ionization is shown to have its maximum value of about 10^6 electrons per cc. at an altitude of about 130 km. by day and 2×10^5 electrons per cc. at 155 km. at night, and to have its lower boundary at an altitude of 90–100 km. (mean values for summer) This ionization can just be produced by the energy of the ultra-violet in sunlight,

and the total conductivity due to this ionization will probably agree fairly well with the value determined from the daily variations of terrestrial magnetism, if Chapman's theory of this phenomenon is altered so as to take into consideration the much greater conductivity of the air in the direction of the magnetic field than perpendicular to it.

On the other hand, starting from this state of ionization all the well-known facts of radio wave propagation may be explained, at least qualitatively and in many cases even quantitatively. The outlines of the process of radio propagation are in general about as follows:

Long waves will be refracted without severe losses from a height of about 90–100 km. even for great earth angles. Both electrons and ions are effective in bending the rays. The direct earth wave is also of importance for distances up to about 1,000 km. and (during summer days) gives the interference pattern found by Hollingworth. At greater distance the refracted waves will predominate; still, the attenuation will not be so great as that found by Zenneck for propagation parallel to the earth because the multiple reflexions will distribute the energy almost uniformly over the space between the earth and the ionized "layer," while according to Zenneck the energy is propagated along the boundary surfaces and is nearly zero somewhere between them. The total energy is therefore actually much greater than that found by Zenneck, while the losses in the earth and in the ionized air are almost the same. Taking this into consideration, the author finds an attenuation for long waves which is in rather good agreement with the practical data as contained in Austin-Cohen's formula.

During summer days, the long waves will be refracted from a rather low height where the air pressure is so high and the number of heavy ions relatively so great that the magnetic field of the earth will have very little influence; the conditions of transmission will therefore be good and steady. During winter days, and during night time, the heights of refraction will be greater so that the magnetic field will be capable of rotating the plane of polarization and giving double refraction, and therefore will cause extremely variable conditions and poor (inaccurate) direction finding. This fact supplies a very good means for determining height of refraction of the very long waves.

For very short wavelengths the ground waves will be absorbed at very small distances from the transmitter. The rays will be

refracted at heights from 130 to 150 km. and by electrons only. Rays of wavelengths shorter than a certain limiting value cannot be refracted back to the earth; these limiting values are assumed to be 8.5 m. by day and 18.9 m. by night. These data determine the maximum numbers of electrons per cc. by day and by night and also a maximum value of the recombination constant at these heights. When the wavelength increases from these limiting values, the electrons will be able to refract a greater part of the radiated energy but at the same time they will give greater losses. For waves having a great part of their paths in or near to the height of maximum electron density there will therefore be a maximum of range for wavelengths of about 15–22 m. These waves are able to encircle the earth, propagated mostly at a height where their radius of curvature is nearly equal to the radius of the earth. When phase- and group-velocities are not confounded, the measurements of the time required by the signals to encircle the earth give average heights of transmission of the right order of magnitude, viz., about 150 km. The fact that it is just possible for a ray of 15 to 22 m. wavelength to be transmitted around the earth gives us rather good values of the number of collisions suffered by an electron per second in these heights and therefore also of the air pressure and assures us that no appreciable amounts of hydrogen can be present there.

Short waves of wavelengths greater than about 22 m. will be refracted from heights of 110 to 140 km. They are best transmitted during night time while the very short waves can only be bent down during day time, when the ionization is maximum. The "skip" phenomena are simply deduced from the shape of the curves of the refraction index and the fact that there is no ground wave except very near to the transmitter.

For waves of medium wavelength the refracted rays will be more attenuated than those of short wavelength and the ground wave will suffer a greater loss than that of a long wave. Therefore the day and night ranges will have minima for wavelengths of about 200 m. The magnetic field of the earth makes the drop even more pronounced, even if there is no real resonance effect because of the collisions between electrons and molecules and because of the influence of the heavy ions.

Those readers who are primarily interested in the practical results will find it most convenient to skip over the mathematical derivations in the first part of the book. They are advised to

start with chapter XI where the author's chief ideas of radio transmission are set forth and to go back to the earlier chapters only now and then for reference.

The book contains a comprehensive set of references to the literature of this field which is of great value in view of the huge number of papers presented on these subjects—in fact, it is hardly feasible to read even the more important of them.

And this bibliography in connection with the profound criticism and the valuable new ideas given in the book—which may really be said to supply us with the missing link between radio experiments and the corresponding fields of physics—strongly supports the hope expressed by the author that it will serve as a basis for the future development both of the physics of the higher atmosphere and of the theory of radio wave propagation.

J. C. F. Rybner

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED
January 4, 1928

Transferred to the Fellow grade

New York	New York City, 434 E. 141st St.	Fried, J. D.

Transferred to the Member grade

New Jersey	Maplewood, 5 Park Ave.	Cunningham, F. W.
Canada	Toronto, 62 Pleasant Blvd.	Patience, A. M.

Elected to the Member grade

California	San Francisco, 611 Crocker Bldg.	Lippincott, D. K.
New York	Brooklyn, 70 Washington St.	Siegil, Samuel
Holland	The Hague, Van Swietenstraat 16	Schotel, Gerrit

Elected to the Associate grade

California	Glendale, 1246 Elm Avenue	Minkler, L. D.
	Los Angeles, 703 W. 81st St.	Dalton, S. P.
	Sacramento, P. O. Box 632	Barlow, Wm.
	San Bernardino, 831 Fifth St.	Beeson, K. H.
	San Diego, 1332 Bush Street	Clement, J. D.
	San Diego, 4657 Oregon St.	Partch, Paul C.
	San Francisco, 812 Hobart Bldg.	Tuel, A. Y.
	San Francisco, 719 Scott St.	Patron, A. E.
Connecticut	Bridgeport, 200 Hazelwood Ave.	Mapes, H. T.
	Waterbury, 9 Burton St.	Greenblatt, Louis
Dist. of Columbia	Washington, Radio Bldg.	Hall, E. F.
Florida	St. Petersburg, 1127—12th St.	Wittstruck, H. H.
Illinois	Bloomington, 511 N. Mason St.	Goodfellow, E. D.
	Chicago, 349 East 59th St.	Cook, B. L.
	Chicago ,2600 Indiana Ave.	Mac Dermain L. D
	Chicago, 4912½ Jackson Blvd.	Ewald, F. J., Jr.
	Chicago, 2011 Jarvis Ave.	Barlow, Wm. H.
	Chicago, 1815 Morse Ave.	John, K. B.
	Chicago, 940 West 55th St.	Noonan, J. J
Indiana	Bloomington, 1000 South Lincoln St.	Hazel, H. C.
Kansas	Topeka, 230 Woodlawn Ave.	Jones, G. I.
Louisiana	New Orleans, 220 So. Johnson St.	Collins, J. F.
Maryland	Baltimore, 530 Richwood Ave.	Houston, G. P.
Massachusetts	Boston, 191 Friend St.	Krikorian, D. K.
	East Springfield, 94 Edendale St.	Kinsman, W. D.
	Malden, 21 Garland Ave.	Winchell, W. H.
Michigan	Detroit, 2914 Book-Cadillac Hotel	Harper, J. K.
	Detroit, 3026 Lothrop Ave.	Byerlay, H. L.
	Detroit, 8037 Medbury Ave.	Friedenthal, A.
	Detroit, 2170 St. Clair Ave.	Ballbach, N. C.
Nebraska	Omaha, 2230 Jones St.	Cramer, W. R.
New Jersey	East Orange, 402 N. Grove St.	Beidleman, H. B.
	Ft. Monmouth, 15th Signal Ser. Co.	Wilson, L. E.
	Glen Ridge, 30 Edgewood Road	Cunliffe, P. R.
	Ridgefield Park, 205 Preston St.	Pearce, Arthur
	West Orange, 7 Helen Avenue	Tucker, M. E.
New York	Albany, 241 Quail St.	Ainsworth, R. I.
	Bronx, 494 Concord Ave.	Zysman, Harry
	Brooklyn, 485 East 21st St.	Tucker, R. A.
	Brooklyn, 755 Gravesend Ave.	Kennedy, H. E.
	Brooklyn, 85 Montrose Ave.	Consoli, Vincent
	Buffalo, 555 East Amherst St.	Stevens, H. A.
	L. I., Elmhurst, 3714—92nd St.	Banks, A. C.
	L. I., Jamaica, 150-08—116th Drive	Albert, M. M.
	New York City, 2022—233 Broadway	Fisk, Wm. G.
	New York City, 771 East 180th St.	Pleasanton, A. W.
	New York City, 463 West St.	Delimel, R. C.
	New York City, 463 West St.	Freeman, R. H.
	New York City, 254 West 98th St.	Kanely, N. J.
	Orient	Vail, L. G.
	Schenectady, 1719 Broadway	Billins, J. H.
	Schenectady, 501 Campbell Ave.	Hoard, N. F.
	Schenectady, 318 Clinton St.	Selby, E. O.
	Schenectady, 143 Division St.	Dodge, A. O.
	Schenectady, 37 Eleanor St.	Stone, F. B.
	Schenectady, 806 Grant Ave.	Hoffer, G. M.
	Schenectady, 9 Park Place	Priebe, Frank
	Schenectady, 606 Pleasant St.	Nickle, E. G.
	Schenectady, 205 Seward Place	Frank, C. W.

	Schenectady, 231 Seward Place	**Newbold, W. H.**
	Schenectady, 929 State St.	**Demmer, A. H.**
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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 16 — March, 1928 — Number 3

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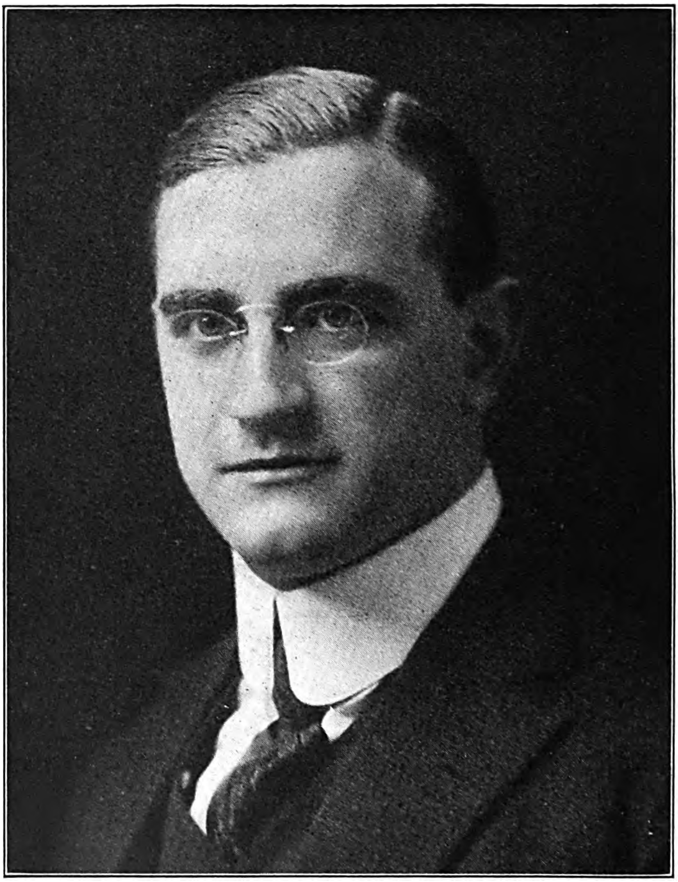


ALFRED N. GOLDSMITH

President of the Institute, 1928

Alfred N. Goldsmith

PRESIDENT OF THE INSTITUTE, 1928

Alfred N. Goldsmith was born in New York City on September 15, 1887. He attended grammar and high school in New York, received the B. S. degree from the College of the City of New York in 1907, and the Ph. D. degree from Columbia University in 1911.

For a number of years he was Associate Professor and Professor in charge of electrical engineering at the College of the City of New York. During 1912 he served as a consulting radio expert with the U. S. Department of Justice and in 1914 as a consulting radio engineer for the Atlantic Communication Company. From 1915 to 1917 Dr. Goldsmith did a considerable amount of consulting engineering work for the General Electric Company until he became director of research of the Marconi Wireless Telegraph Company of America. Upon the formation of the Radio Corporation of America he was appointed Director of Research. Since 1923 he has been Chief Broadcast Engineer of the Radio Corporation of America and in 1927 became Chairman of the Board of Consulting Engineers of the National Broadcasting Company.

During the World War he was Technical Director of the U. S. Signal Corps School of Communication and the U. S. Naval Radio School at the College of the City of New York.

To most members of the Institute Dr. Goldsmith is best known for his continuously active service for the Institute since its inception in 1912 He was one of the three or four engineers responsible for the organization of the Institute and has served continuously as Editor of the Institute's publications since 1913. From 1918 to 1928 Dr. Goldsmith was Secretary of the Institute. He has been a member of the Board of Direction of the Institute since 1913, and at one time or another has served in practically all capacities on the Institute's various committees.

Dr. Goldsmith has contributed a number of technical papers to the PROCEEDINGS of the Institute. He has been a Fellow in the Institute since 1915.

CONTRIBUTORS TO THIS ISSUE

Anderson, Clifford N.: Born at Scandinavia, Wisconsin, September 22, 1895. Graduated from Wisconsin State Normal School in 1913. From 1913 to 1917, supervising Principal of Schools, Amery, Wisconsin. During the World War served as Ensign, U.S.N.R.F. in the Aviation Section; received the Ph. B. degree from the University of Wisconsin in 1919 and the M. S. degree in 1920. During 1919 and 1920 served as Instructor in Engineering Physics at the University of Wisconsin; was with General Electric Company, Lynn, Mass. from 1920 to 1921; Fellow, American Scandinavian Foundation to Norway 1921–1922; from 1922 to date with the American Telephone and Telegraph Company, Department of Development and Research doing work in connection with transatlantic radio telephony. Associate member of the Institute.

Austin, L. W.: See PROCEEDINGS for February, 1928.

Bowditch, F. T.: Received the B. S. degree in electrical engineering from the University of Illinois in 1919. Was in the Research Laboratory of the Aluminum Castings Company (now a subsidiary of the Aluminum Company of America), 1919–1920. From 1920 to date has been a radio engineer in the Research Laboratory of the National Carbon Company. Associate member of the Institute.

Dahl, Odd: Born at Drammen, Norway, November 3, 1898. Introduced short-distance radio telephones in the Norwegian fishing industry in 1919; lieutenant in the Norwegian Army Air Force, 1920; aviator and radio engineer with Captain Roald Amundsen's *Maud* Expedition, 1922–1925; since January 1927, assistant physicist, Department of Terrestrial Magnetism, Carnegie Institution of Washington.

Gebhard, Louis A.: Born at Buffalo, New York, June 11, 1896. Received the LL. B. degree at Georgetown University; commercial radio operator, 1913–1917. In the naval radio service 1917–1923 at the Naval Radio Laboratory, Great Lakes, Ill.; Naval Radio Station, Belmar, N. J.; Naval Aircraft Radio Laboratory, Naval Air Station, Anacostia, D. C. Since 1923 at Naval Research Laboratory, Bellevue, Anacostia, D. C. At present in charge of the high-power transmitter department of the Naval Research Laboratory. Associate member of the Institute.

Hulburt, E. O.: See PROCEEDINGS for February, 1928.

Loftin, Edward H.: Born July 19, 1885 at Montgomery, Alabama. Graduated, U. S. Naval Academy, 1908; post graduate course at Naval Post Graduate School, Annapolis and Columbia University with M. A. degree; closely affiliated with naval radio activities from 1910 to 1924 during which time commanded naval radio research ship *Bailey*, pioneered development for radio aircraft, in charge of naval communications in France during the War, liaison officer on Inter-Allied Conferences, negotiations for and construction of the Lafayette station in France, in charge of naval research and development of radio for four years after war, member of Technical Committee of International Communication Conference (1921), and chairman of Inter-Departmental Radio Board. Since leaving the naval service in 1924

has been engaged in private research and development work. Member of the Institute.

Walmsley, Thomas: Born Burnley, Lancs., England in 1886. Graduate of Leeds Grammar School, Leeds Technical School, London Polytechnics School with the B. Sc. degree in 1917. From 1908 to date has held various positions in the British Post Office Engineering Department in both Central Power Section and Wireless Section; from 1920 to 1922 was engineer in charge of Abu Zabaal (Cairo) Wireless Station; since 1923 has been engineer in charge of the Rugby Radio Station GBR. Member of the Institute.

White, S. Young: Born April 11, 1901. For some time in testing course of the General Electric Company; spent a number of years as radio operator on ships throughout the world; engaged in special work on rectifiers in the United States and Europe for several years; during the past few years has been engaged in private research and development work, particularly in amplifying circuit and radio design work.

Wright, C. A.: Received the B. S. degree in M. E. and E. E., Tulane University, 1906; U. S. Engineer Office, Vicksburg, Miss., 1907; Testing Department, General Electric Company, Schenectady, N. Y., 1907–1908; Instructor in electrical engineering, graduate school of applied sciences, Harvard University, 1908–1909; received the E. E. degree (in absentia) from Tulane University in 1909; 1910–1915, telephone engineer with the American Telephone and Telegraph Company; Associate Professor of electrical engineering, Iowa State College, 1915–1918; 1918–1926, Professor of electrical engineering, Ohio State University; 1926 to date, Radio Engineer, Research Laboratory, National Carbon Company, Inc. Associate member of the Institute.

INSTITUTE ACTIVITIES

February Meeting of the Board of Direction

At the February meeting of the Board of Direction, held on February 1st in the offices of the Institute, the following were present: Alfred N. Goldsmith, President; L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Ralph Bown and Donald McNicol, Junior Past Presidents; Arthur Batcheller, W. G. Cady, R. A. Heising, R. H. Manson, R. H. Marriott, and J. M. Clayton, Acting Secretary.

Dr. Goldsmith was re-elected Editor of Publications, and Mr. Eastham, Treasurer. J. M. Clayton was elected Secretary. Dr. Leonard F. Fuller was elected a Manager of the Institute.

The following transfers and elections, upon recommendation of the Committee on Admissions, were authorized: Transfer to the grade of Member: H. A. G. Howse, E. M. Dupree, and C. M. Srebroff; Election to the grade of Member: E. A. Lederer, D. V. Adendorff, G. E. Bliziotis, D. H. Newman, M. L. Prescott, K. E. Rollefson, G. E. J. Suadicani, and W. A. Tolson.

One hundred and fifteen Associate members and eighteen Junior members were elected.

Patent Digests

The Board of Direction requests certain information from the Institute membership as to the continued publication of the Patent Digest pages of the Proceedings. The Board feels that these patents, as they have been published during the past several years, do not usefully serve the membership. Several questions are addressed to the membership on page 388 of this issue. The Board will be guided by the number and character of replies received to these questions. If the Patent Digests are discontinued, several additional pages of papers can be printed each month. If the Digests are to be resumed in any of the forms outlined on page 388, it may be necessary to decrease the number of editorial pages by an amount depending upon the space occupied by the Digests. Your comments are desired immediately.

Changes of Address

Members of the Institute are again cautioned to advise the office of the Institute in the event of any change in their mailing address in order to insure prompt and uninterrupted receipt of all

Institute publications. The Institute cannot be held responsible for failure to receive copies of the PROCEEDINGS if it has not been notified of changes in mailing addresses.

Committee Work

PROGRESS REPORT OF COMMITTEE ON STANDARDIZATION, 1927

The following report of the activities and progress of the Institute's Committee on Standardization has been submitted by the Chairman of the Committee, L. E. Whittemore:

The work of the Committee on Standardization has been carried on during 1927, as in 1926, largely through five subcommittees. The chairmen of these subcommittees during the year 1927 were as follows:

1. Vacuum Tubes, L. A. Hazeltine;
2. Circuit Elements, H. M. Turner;
3. Receiving Sets, J. H. Dellinger;
4. Electro-Acoustic Devices, R. H. Manson;
5. Power Supply, Walter E. Holland.

Preliminary drafts of reports of three of these subcommittees, namely, Vacuum Tubes, Receiving Sets, and Electro-Acoustic Devices, were printed under date of May 20, 1927, and were given fairly wide distribution in order that criticism and comments might be secured before their final consideration by the Committee on Standardization. In general, these subcommittee reports covered the formulation of methods expressing and measuring the characteristics of radio apparatus and devices falling within their respective fields. No general effort is being made by the subcommittees to revise the definitions contained in the Standardization Report published in 1926, but recommendations are being made as to new definitions or symbols, or changes in the 1926 definitions which the subcommittees find during the course of their work to be considerable.

During the course of the year there has been some discussion of the relation between the standardization work of the Institute of Radio Engineers and that of the associations of radio manufacturers. The chairman of the I. R. E. Standardization Committee has had some correspondence with a number of persons interested in this question and has found that the following statement appears to meet with general acceptance as an expression of the relationship between the standardization work of these two groups. It

is recognized, of course, that in this new field it is impossible at the present time to determine upon any hard and fast dividing line.

Institute of Radio Engineers

(1) terms, definitions and symbols, and

(2) methods of testing materials and apparatus in order to determine their important characteristics. This work may consist of purely advisory discussion as to convenient forms of tests, precautions to be taken, etc., or it may include standardization of definite test procedures to serve as a common basis of comparison of the properties or performance of material or apparatus.

Manufacturers' Groups

(1) standardization of size and characteristics of apparatus, to promote interchangeability of parts, either mechanical or electrical, and

(2) setting of standard ratings for the properties or performance of material or apparatus.

At the request of the Federal Radio Commission a study was made of the suggestion that a system of assignments of frequencies to broadcasting stations be employed in which the frequencies would be those ending in the digit "5" rather than the digit "10", as at present. After correspondence with the members of the Standardization Committee, the Board of Direction and others, a report was prepared entitled "Discussion of Suggested Use of Broadcasting Frequencies Ending with the Digit 5." This report, dated May 23, 1927, was transmitted by President Bown of the Institute to the Federal Radio Commission.

Also at the request of the Federal Radio Commission, the Committee has been making a study of the power rating of radio stations. A draft of a report on this subject is in preparation. The completion of this report is awaiting the result of some additional correspondence with the Commission.

On several occasions during the work of the subcommittees the suggestion has been made that a special subcommittee be appointed to study the use of the transmission unit as an expression of the characteristics of radio circuits and devices. The subcommittee on Electro-Acoustic Devices, in particular, urged that this be done. A subcommittee on this subject was therefore appointed, and J. V. L. Hogan has consented to serve as its chairman.

The Subcommittee on Receiving Sets has recommended the appointment of a special subcommittee to prepare a bibliography covering the fields of five of the principal subcommittees. Bibliographical material has been prepared by the Subcommittee on Receiving Sets and the Subcommittee on Circuit Elements, and it is recommended that similar bibliographies covering the subjects of the other subcommittees be prepared for publication with the final report.

It is believed that the several subcommittees can complete their work during the first two or three months of 1928 and that a complete report of the Committee on Standardization can be published late in the spring of this year.

Professor Hazeltine served as acting Chairman of the Committee on Standardization during the last half of 1927 during the absence of the Chairman.

The progress made by the several subcommittees during the year 1927 may be summarized briefly as follows:

1. Vacuum Tubes—L. A. Hazeltine

A final draft of a report containing some modifications and additions to the preliminary draft of May 20, 1927, was circulated to members of the subcommittees under date of January 3, 1928. A few minor changes may be required but the report is believed to be substantially ready for final action by the main committee.

The symposium on the measurement of the inter-electrode capacity of vacuum tubes, forming a part of the 1928 I. R. E. convention program, was a direct outcome of the work of the members of this committee and those who are associated with them.

2. Circuit Elements—H. M. Turner

Material has been collected for a report of this subcommittee. A chart has been prepared giving references to several methods which are available for measuring the various properties of radio-circuit elements, and a bibliography has been compiled. It is expected that a report of this subcommittee will be prepared during the spring of 1928.

3. Receiving Sets—J. H. Dellinger

Meetings of this subcommittee were held on March 3, May 5, November 1, 1927. The preliminary draft report of May 20, 1927

has been revised and is being extended through the work of four subcommittees on the following subjects:

1. Field intensity measuring apparatus.
2. Test procedures.
3. Bibliography.
4. Additional quantities (additional to sensitivity, selectivity, and fidelity).

It is believed that the work of this subcommittee will lead to a revised report in the spring of 1928. This will not be a final report as the work of the subcommittee on this subject is likely to extend over several years.

4. Electro-Acoustic Devices—R. H. Manson

Several subcommittees have been working on special questions on which they expect to report at a meeting to be held early in February. At this time the preliminary draft of May 20 will be improved and probably somewhat extended. It is anticipated that a final report of this subcommittee will be ready some time in March.

5. Power Supply—W. E. Holland

This subcommittee has reviewed the present I. R. E. standard terms which pertain to power supply. It has suggested revisions in some of the definitions and has originated such new terms and definitions as seem to be needed. This work has been done in several meetings, one of which was a joint meeting with the Socket Power Committee of the Radio Division of the National Electrical Manufacturers' Association.

This subcommittee also has been making a study of the subjects listed below, and it anticipates that reports will be ready in the near future on one or more of these subjects.

(1) Methods of measuring inductances of iron core inductors (in coordination with the Subcommittee on Circuit Elements).

(2) Method of measuring capacitance and power factor of filter capacitors.

(3) Methods of measuring ripple in the output of power supply devices.

To all members of the Committee on Standardization for 1927 and its Subcommittees and especially to the chairmen of these subcommittees, the members of the Institute owe their thanks for the cooperation and splendid service given during the past year.

Institute Meetings

New York Meeting

At the New York meeting of the Institute held on the evening of February 1st in the Engineering Societies Building, two papers were delivered. The first, by Frederick E. Terman of Stanford University, entitled "The Inverted Vacuum Tube, A Voltage-Reducing Power Amplifier," was read by Dr. A. N. Goldsmith. In the discussion which followed the reading of the paper the following, among others, took part: Alfred N. Goldsmith, George Crom, and Harold A. Wheeler.

The second paper of the evening was presented by Professor Hidetsugu Yagi, of Tohoku Imperial University, Japan. The subject was "Beams of Ultra Short Waves." Messrs. Goldsmith, Kroger, Yagi, Crom, Martin, Whittemore, Hallborg, Bohn and others discussed the paper.

Both papers were profusely illustrated with lantern slides. They will be published in a forthcoming issue of the Proceedings.

The attendance at this meeting was about two hundred and seventy-five.

Atlanta Section

A meeting of the Atlanta Section was held on February 1, 1928 in the Chamber of Commerce Building, Atlanta, Georgia. Major Walter Van Nostrand, Chairman of the Section, presented a paper on "Advancements in the Measurement of Radio Frequencies."

The next meeting of the Atlanta Section will be held in the Chamber of Commerce Building on March 7th.

Canadian Section

On January 19, 1928 a meeting of the Canadian Section was held in Room 23 of the Electrical Building, University of Toronto. A. M. Patience presided.

C. V. Loughren, of the Research Laboratory of the General Electric Company, presented a paper on "A-C. Vacuum Tubes."

Messrs. Bailey, Baldwin, Price, Northover, Hepburn, and others discussed the paper.

The fifth lecture of the Junior Lectures on "Vacuum Tubes" was delivered by C. C. Meredith. Messrs. Price, Loughren, Lowry, Smith, Baldwin, and others discussed the lecture.

It was announced that the winners of the Membership Competition were (1) M. Hodsoll and (2) C. I. Soucy.

Eighty members attended the meeting.

On February 1st the Canadian Section held a meeting in the Electrical Building of the University of Toronto. Two papers were presented. The first, by G. E. Pipe of the Standard Radio Company, was entitled "Radio-Frequency Circuits." C. I. Soucy and others discussed this paper.

The second paper of the evening was on "Uni-Control Receivers," and was presented by Walter Jones of the Federal Radio Corporation of Buffalo. This paper was discussed by Messrs. Bagley, Soucy, Patience and others.

Sixty-one members attended this session.

The next meeting of the Section will be held on March 14th at which time R. C. Hitchcock, of the Westinghouse Company of Pittsburgh, will deliver a paper on "Crystal Control." The meeting will be held in the Electrical Building of the University of Toronto.

Cleveland Section

In the Case School of Applied Science, Cleveland, a meeting of the Cleveland Section was held on February 3rd. John R. Martin presided.

A paper by Drs. P. L. Hoover and C. Nusbaum, of the Case School of Applied Science, on "The Design of Filter Circuits" was the first paper of the evening.

The second, by Dr. C. Nusbaum, was entitled "Emission from Hot Bodies."

Seventy-five members and guests attended the meeting.

Detroit Section

The Detroit Section held a meeting on January 20th in the Conference Room of the Detroit News Building. Thomas E. Clark presided.

Captain Norman L. Baldwin, Signal Corps, U. S. Army, delivered a paper on "Short-Wave Transmission and Reception." A general discussion followed the presentation of the paper.

Officers of the Section for 1928 were elected as follows: Earle D. Glatzel, Chairman; A. B. Buchanan, Vice-Chairman; Walter R. Hoffman, Secretary-Treasurer. L. N. Holland was appointed Chairman of the Meetings and Papers Committee, and Charles H. Cox was made Chairman of the Membership Committee.

The next meeting of the Section will be held on February 17, 1928 in the Engineering Building, University of Michigan, Ann Arbor.

Philadelphia Section

On January 27th a meeting of the Philadelphia Section was held in the Franklin Institute. J. C. Van Horn presided.

The speaker of the evening was Carl Dreher of the National Broadcasting Company, who delivered a paper on "Broadcast Operation". Messrs. Frazier, Miller, and others discussed the paper.

Sixty members and guests attended the meeting.

The next meeting will be held on February 24th in Franklin Institute.

Rochester Section

A meeting of the Rochester Section was held in the Sagamore Hotel on January 13, 1928. Joseph Hichcock presided.

George B. Crouse, of the Conner-Crouse Corporation, presented a paper on "Development of Line 'A' Power."

Sixty members attended the meeting.

On February 3, 1928 another meeting of the Section was held in the Sagamore Hotel. H. J. Klumb presided.

A. Acheson, of the General Electric Company, presented a paper entitled "Broadcasting One Hundred Thousand Watts." The paper included lantern slides and the display of a 100,000-watt tube.

On March 2nd Hugh M. Stoller, of the Bell Telephone Laboratories, will present a paper on "Television" in the Hotel Sagamore at a regular meeting of the Section.

San Francisco Section

Upon the completion of the reorganization of the San Francisco Section the following officers for 1928 have been elected: Chairman, Dr. Leonard F. Fuller; Vice-Chairman, Donald Lippincott; Secretary, D. B. McGown.

On January 24, 1928 the first meeting of the year of the San Francisco Section was held. This was a joint meeting between the Institute of Radio Engineers San Francisco Section and the local American Institute of Electrical Engineers Section.

Dr. E. B. Craft, of the Bell Telephone Laboratories, delivered a paper on "Coordination of Research." Each person in attendance

received a book showing the laboratory activities of the laboratories of the Bell Company.

Over four hundred and fifty persons attended the meeting. An informal dinner preceding the meeting was held in the Club Rooms of the San Francisco Engineers Club. One hundred and six members and guests attended the dinner.

Seattle Section

The Seattle Section of the Institute held a meeting on January 28, 1928 in the Club Rooms of the Telephone Building, Seattle. T. M. Libby presided.

E. L. White presented a paper, "Location and Mitigation of Radio Interference in Broadcast Receivers," and E. H. Schrieber presented a paper, "General Aspects of Radio Interference."

Messrs. Libby, Wilson, Burleigh, Kleist, and others discussed the papers.

The result of the election of 1928 officers of the Seattle Section is as follows:

Chairman, W. A. Kleist; Secretary-Treasurer, Oliver C. Smith. The following standing committees were appointed by the Chairman: Meetings and Papers: J. R. Tolmie (Chairman), A. V. Eastman, C. E. Williams and H. F. Mason. Membership Committee: J. A. Burleigh, H. E. Renfro, T. M. Libby, and W. A. Douglass.

Washington Section

On February 9, 1928 in Picardi's Cafe, 1417 New York Avenue, N. W., a meeting of the Washington Section was held. A. Hoyt Taylor presided.

Professor Hidetsugu Yagi, of Japan, presented a paper, "Beams of Ultra Short Waves," which was profusely illustrated with lantern slides. Following the presentation of the paper the following took part in the discussion: A. Hoyt Taylor, J. H. Dellinger, August Hund, Major Blair, Mr. Robinson, and others.

Dr. Dellinger spoke briefly on the New York Convention of the Institute and on the award of the Morris Liebmann Memorial Prize to Dr. Taylor.

The next meeting of the Washington Section will be held on March 8, 1928 at which time Dr. A. Hoyt Taylor and L. C. Young, of the Naval Research Laboratory, will deliver a paper on "Sending and Receiving Radio Signals Around the World."

Over seventy-five members attended the meeting.

Personal Mention

C. L. Davis, of Milwaukee, Wisconsin, has joined the staff of John B. Brady, Patent Attorney, of Washington, D. C.

Dr. M. A. F. Barnett, who has been at Clare College, Cambridge, England, is now Physicist at the Dominion Laboratory, Wellington, New Zealand.

Bernard H. Linden, formerly Radio Inspector in the Sixth Inspection District, has been promoted to the position of Supervisor of Radio with Headquarters at San Francisco.

H. T. Melhuish, late manager of sales administration of the Radio Corporation of America, has joined the staff of the National Electrical Manufacturers' Association as Director of the Radio Division.

Charles V. Litton has joined the engineering staff of the Federal Telegraph Company of Palo Alto, California. Mr. Litton was with the Bell Telephone Laboratories, stationed at Deal, New Jersey.

Alan N. Ramsay recently resigned as Sales Manager of Sherman, Clay and Company of Los Angeles to become General Sales Manager for the Precision Electric Manufacturing Corporation of the same city.

ENGINEERING SOCIETIES BUILDING

New Headquarters of the Institute of Radio Engineers

New Quarters of the Institute

The offices of the Institute have been removed from 37 West 39th Street, New York, to the 8th floor of the Engineering Societies Building, 33 West 39th Street.

To provide for the increased office staff at Headquarters of the Institute, approximately fifteen hundred square feet of floor space have been secured.

The Headquarters now contain adequate space to carry on the many duties involved in the operation of the Institute and the publication of the PROCEEDINGS.

In addition a Members' Room for the use of members of the Institute has been provided. This room contains facilities for correspondence, copies of current radio periodicals, and provides a place in which members of the Institute may meet.

Members of the Institute are invited to partake of the facilities offered in the Engineering Societies Library located on the thirteenth floor of the Engineering Societies Building. This library contains a collection of over one hundred thousand technical books, and several hundred monthly engineering journals. Photostat copies of pages from any book in the Library can be obtained very reasonably. The Library is also prepared to furnish information monthly as to the publication of papers and articles appearing in current journals and engineering magazines, both domestic and foreign, at very reasonable prices. The library is open every day except Sunday and can be used by all members of the Institute. Members residing out of town are invited to communicate either with the Institute or directly with the Library for information regarding a number of services which can be rendered to them.

The regular monthly New York meetings of the Institute have been held in the meeting rooms in the Engineering Societies Building for a number of years. It is a source of much gratification to the Board of Direction of the Institute that our Headquarters are now located in this splendid building, a photograph of which appears on the opposite page.

Members visiting New York are invited to make The Institute Offices their Headquarters for their correspondence and meeting place.

OBITUARY

With deepest regret the Institute announces the death of

Robert Loughry

Colonel Loughry served continuously in the Army of the United States from the beginning of the Spanish-American to the close of the World War. During the World War he served as Lieutenant Colonel in charge of all Army radio communication on the American front. For his skillful organization of the wartime radio activities he was cited by General Pershing both as a radio officer of the First American Army and later as radio officer of the Zone of Armies.

Since 1922 Colonel Loughry had been Radio Engineer of the 9th Corps Area in San Francisco. At the time of his death he was Commander of the California and Nevada Department of the Veterans of Foreign Wars.

His great genius for organization, and his infectious enthusiasm and unlimited capacity for hard work were surpassed by his kindly, genial nature and quality of friendliness, rarely found in men of his ability. For these he was best loved by his comrades.

Colonel Loughry was a Fellow in the Institute.

OPENING ADDRESS OF RALPH BOWN RETIRING PRESIDENT

Annual Meeting and Convention, Institute of Radio Engineers, January 9, 1928

At this the annual meeting of the Institute it is customary that the retiring president give a summary of the activities of the organization. I find upon looking over again the report given by Mr. McNicol at the last annual meeting that much of historical importance was included in his remarks which became of record* and therefore does not need repetition.

The time required for the registration of so large a number of members as is present today has already cut heavily into the time available for this session and there is need for brevity in its proceedings. I shall therefore merely touch upon a few of the more important matters of the past year.

On January 1, 1927 the Institute entered upon a new phase of its development. Employee management of its office and many of its business affairs was initiated. A year's trial of this new departure has proved it to have been a wise one. The Institute has prospered. Many things hopefully planned in previous years it has been possible to bring to fruition. It seems appropriate to mention in this connection that our Assistant Secretary, Mr. John M. Clayton, deserves major credit for the success of the new policy.

At the beginning of the year the actual paid membership was approximately 2800. There were also being carried on the lists nearly a thousand names of members who had become inactive and which have since been deleted. The growth in new membership has been large and steady; the paid membership has now reached over 4200, an increase of about 55 per cent during the year.

The increase in income from dues represents a considerable increase in the margin between the basic running expenses and the total income. This margin, except for a small surplus to be carried into the protective fund, has been spent in ways which should carry it back directly to the membership. The number of issues of the Proceedings has been increased from six to twelve a year and 34 per cent more technical material published. Preprints of papers were provided on a much better scale than had

* See Proceedings for February, 1927.

been possible before. A complete year book was sent to each member and preliminary reports of standardization work were published and circulated. A fourteen-year index of the PROCEEDINGS has been furnished each member.

Without going through the list to mention the particular achievements of each, it may be said that all of the standing committees have shared in the progress of the year. The increased membership, secured in part through good publicity; the standardization reports; the expansion in publications; the new Sections; all reflect credit on the work of the standing committees and their chairmen.

The future growth of the Institute is largely bound up in the fortunes of its Sections. Four new Sections were organized in 1927, at Atlanta, Buffalo, Cleveland, and Detroit respectively. The PROCEEDINGS furnishes the primary bond between the Institute and its members. But membership in the Institute, to be an attractive and useful thing of real value to engineers, must mean more than subscription to a technical periodical even though it be a periodical of great merit. It must mean frequent opportunity to attend meetings for the reading and discussion of technical papers, personal contact with other engineers of the community, and opportunity to hear and meet engineers from other communities. These aspects of the Institute, so familiar to members in the metropolitan district, can be furnished to members in other parts of the country only by active, well-organized Sections. I was fortunate enough last Fall to have been invited to talk at meetings of five different Sections in the eastern part of the United States. To visit these Sections and participate in their meetings was a pleasant and profitable experience and it left with me a firm conviction that the section idea is a sound basis upon which to build a better and larger Institute which has a personal value to each member.

The Institute Gold Medal for 1927 was presented at the June meeting to Dr. L. W. Austin for his pioneer work in the quantitative measurement of radio transmission. It was noted at that time that the study of radio transmission, in the initiation of which Dr. Austin had so important a part, has now become one of the most engaging present-day activities of radio engineers. The apparatus side of our art has to a degree outrun the transmission side and we are now endeavoring that it catch up.

It is now my pleasant duty to represent the Institute in the bestowal of the Morris Liebmann Prize for 1927. The meritorious work which this prize recognizes was done in the very newest field of radio transmission study, the field of short waves. It seems entirely appropriate that its recipient should be a member of the United States Navy, which has been so forward-looking in the application and use of this novel mode of sending human intelligence through space. Whether Dr. Taylor coined the term "skip distance" I do not know, but there is no doubt that he has made very important contributions to our knowledge of this phenomenon. It is a pleasure to present this prize to Dr. A. Hoyt Taylor.

Before relinquishing the duties of office I wish to tell you that I have thoroughly enjoyed being president of our Institute and to thank you for the privilege of serving in that capacity. It has been a beneficial experience to me. I only hope that some small benefit may have accrued to the Institute.

Our new president is so well-known to you that anything in the way of an introduction would be superfluous. He might almost be called the godfather of the Institute, since in his capacity as Secretary and Editor of Publications he held it together through many lean years and contributed much to building the firm foundation on which it now stands. It seems particularly fitting that he should become our president at a time when the Institute is so flourishing. Our president, Dr. Alfred N. Goldsmith.

ADDRESS OF DR. ALFRED N. GOLDSMITH

President of the Institute of Radio Engineers

It is rather overwhelming to contemplate the marvelous strides which have been made by the Institute of Radio Engineers in its fifteen-year growth. Originating among "a little group of serious thinkers" in the radio field who felt the urgent need for some means for self-expression and mutual cooperation among radio engineers, it has grown to be an internationally known and leading engineering association, with many thousands of members, and with a publication, its PROCEEDINGS, which is to be found on the desk of practically every active worker in the radio field.

It is the finest sort of an honor to have been chosen as the President of such an organization, however undeserved may be this friendly commendation from the Institute membership.

High standards have been set by past Presidents of this Institute. Let me recall to you the list of these gentlemen who have worked so effectively and loyally in order that the Institute might be what it is today. In order of their tenure of office, they are:

Mr. R. H. Marriott, veteran radio engineer and indefatigable organizer of the Institute.

Mr. G. W. Pickard, a radio scientist noted for the exquisite elegance of his researches and no less for the continued service he has given the Institute.

Dr. L. W. Austin, who has literally been a right-hand of the Government in radio matters and has shown the way in orderly measurement of complicated radio transmission phenomena.

Mr. John Stone, a pioneer in the wire telephone field, and in radio circuit investigations.

Prof. A. E. Kennelly, a master electro-mathematician who brought all the wisdom of his past presidency of the American Institute of Electrical Engineers to the aid of the Institute of Radio Engineers.

Prof. M. I. Pupin, that brilliant scientist of vivid and charming personality. If I did not fear that it would detract from his otherwise flawless reputation, I would stress the pleasure I experienced while I studied under his able direction.

Prof. G. W. Pierce, one of America's leading electrical teachers and investigators.

Mr. J. V. L. Hogan, a gentleman whose combined engineering skill and knowledge and executive ability have ever been freely at the disposal of the Institute.

Mr. E. F. W. Alexanderson, a most original and fruitful worker in the field of radio.

Dr. Fulton Cutting, who may be described as a "gentleman scientist," in the finest sense, in the field of radio research.

Dr. Irving Langmuir, a research worker combining rare originality and the capability of applying effectively abstruse basic principles.

Prof. J. H. Morecroft, one of the country's best-known radio teachers and a prominent author of radio texts from which we all learn.

Dr. J. H. Dellinger, who may be termed a national and international "liaison" officer in every important aspect of radio.

Dr. Donald McNicol, veteran wire telegraph engineer, and radio worker and organizer; and

Dr. Ralph Bown, one of the leaders in the public service engineering applications of radio and a most capable and effective executive of the Institute.

With these gentlemen as my predecessors, you will readily appreciate that the truly hard work has been largely done and that the credit for reaping what they have sown should be justly given to them.

These gentlemen have carried forward radio development through this Institute during the most romantic period in the radio field. Every communication field has been romantic at one time or another although most of us have had no opportunity to realize the romance of other communication fields than radio.

And yet wire telegraphs and cables were once most romantic. As late as 1885 Owen Meredith in his novel "Lucile" described the plight of a gentleman whose financial fortunes had fallen into sad confusion because there was no way of getting word to him quickly of the impending failure of his banker. Meredith in a quizzical footnote comments poetically on the situation:

> These events, it is needless to say, Mr. Morse,
> Took place when Bad News as yet travel'd by horse;
> Ere the world, like a cockchafer, buzz'd on a wire,
> Or Time was calcined by electrical fire;
> Ere a cable went under the hoary Atlantic,
> Or the word "Telegram" drove grammarians frantic.

That same year, a musical gentleman by the name of H. H. Thiele composed a rather indifferent piece of music which he termed "The Telephone Gallop" which starts picturesquely with a trill which is supposed to represent a telephone ringing, and two emphatic notes which are labeled "Hello!" More than thirty years later, as a result of experimental radio telephony from Arlington to Honolulu, one of a later generation of musicians was moved to write a parallel selection entitled "Hello, Hawaii, How Are You?" Thus does history repeat itself.

The other day I found among my papers a yellowed newspaper clipping dated April 25, 1915. I think it will interest you to hear its contents because it gives you some idea of what the Institute was doing only thirteen years ago. And yet these thirteen years have been so crowded with events that those of us here present who are

described in this clipping are already finding their memories of those days dimming. This is the clipping:

Wireless Men at Dinner

Radio Engineers Entertain for Professors Ferdinand Braun and Johann Zenneck

> Seventy members and guests of the Institute of Radio Engineers, including representatives of the wireless systems of England and Germany, last night gave a dinner at Luchow's to Professors Ferdinand Braun and Johann Zenneck, of the German Telefunken Wireless Co. John S. Stone, president of the Institute, was toastmaster. Besides Professors Braun and Zenneck those at the speakers' table were: Dr. Lee DeForest; Judge Julius M. Mayer; Nikola Tesla; Dr. Fritz Lowenstein; Robert H. Marriott, U. S. Radio Inspector; William Dubilier; Professor Alfred Goldsmith; Roy E. Weagant; Edward J. Nally; and Professor G. W. Pierce. A miniature wireless plant on the speakers' table sent out messages during the dinner.

Today new radio triumphs loom before us, shortly to be achieved. Such startling developments as international broadcasting combining the world into a unit, and television which shall carry the aspect of nature and man to all parts of the habitable globe, seem to be "just around the corner". It is no wonder that this Institute of Radio Engineers has progressed with most unusual rapidity. It is composed of televisionaries—men who see far into the future and are yet realists. May this next year of the Institute's activities be in some measure a worthy continuation of its splendid past history.

ON THE DISTORTIONLESS RECEPTION OF A MODULATED WAVE AND ITS RELATION TO SELECTIVITY*

By

Frederick K. Vreeland

Summary—The importance of overtones in the faithful reproduction of speech and music. Overtones are transmitted by the extreme side bands of a modulated wave. Distortionless reception of the full side bands necessary for faithful reproduction. Crowding of the air channels brings adjacent waves into such close juxta-position that selectivity demands a sharp cut-off at the limits of the band. Selectivity by the usual resonance methods trims the side bands. Attempts at compromise by employing damped resonant circuits do not achieve distortionless reception and sacrifice selectivity. Description of an amplifier giving uniform amplification over the entire band width with a sharp cut-off. Description of a band selector having an approximately rectangular frequency characteristic. Frequency control and volume control. Various applications of the band amplifier and the band selector to broadcast reception.

The Problem of Fidelity in Radio Telephony

It is well known to communication engineers that faithful reproduction of music and clear articulation in speech require distortionless transmission and reception of a wide range of audio frequencies. The vibrations of music, and especially the vibrations of articulate speech, are exceedingly complex. When analyzed, however, it is found that the most intricate sound may be reproduced by the combination of a sufficient number of simple harmonic vibrations of different frequencies, combined in the proper relative intensities. Distortionless transmission means that each of these component vibrations must be represented by an alternating current wave of corresponding frequency and that the relative intensities of these current waves must truly correspond to the relative intensities of the component sound waves. These currents must be transmitted without change in their mutual relations and translated again into sound waves, while preserving their true relative intensities.

The problem of distortionless transmission and reception thus means that each of the component frequencies of the sound must go through the many transformations of the signal energy with equal efficiency and all must come out in the finished product in their true mutual relations.

This is an intricate matter. In the field of wire telephony many years of research, involving investigations and technique of the

*Original Manuscript Received by the Institute, December 23, 1927.

highest order, have produced the perfected results of the present day. Much of this knowledge is applicable in the field of radio telephony, and has been used to good advantage, but radio has its own peculiar problems. At the transmitting end of the radio telephone system the terminal problems are similar to those involved in wire telephony and have been met by similar though specialized means. Between the transmitter and the receiver there is no line to introduce distortion, and the causes of fading distortion appear to be beyond our reach at the present time. To this extent the problem is simplified for the radio engineer. But the receiving end of the system involves peculiar difficulties due to the fact that the transmitted waves are not directed to individual receivers and so must be separated from each other by frequency selection.

Selectivity requires that a certain group of radio frequencies, comprising a modulated signal wave, must be separated from all other radio waves.

Fidelity requires that all the frequencies included in the transmitted wave must be received in their true relative proportions.

These two conditions give rise to a peculiar and highly specialized problem of receiver design, namely to construct a selective system that is equally responsive to all frequencies within a given range or band and unresponsive to frequencies outside this band. And the condition of crowded air channels requires a very sharp cut-off at the edges of the band.

The Importance of Overtones in Sound Reproduction

It becomes important therefore to determine what range of frequencies must be included in sound reproduction to give a faithful and satisfactory result. There has been much discussion on this subject and much difference of opinion. The human ear is a marvelously tolerant organ, and the mind by training is able to supply gross deficiencies in the sound. The tolerance of the human ear to distortion made wire telephony possible during the many years that preceded its present high state of perfection. This tolerance is taxed severely, even at the present day, in radio broadcast reception, but it is now generally recognized that faithful reproduction is highly desirable, and once it is made available it will be demanded by broadcast listeners.

A mental picture of the range of audible frequencies may be obtained by reference to the scale of a piano. This ordinarily includes a little over seven octaves, the lowest note having a frequency of approximately 26 cycles per second and the highest note 4,096 cycles. But this is only half the story. Each note on the piano includes not only the fundamental frequency but also many harmonic overtones. Harmonics at least as high as the eighth, that is, eight times the fundamental frequency, play an important part in determining the tone of the piano, and the seventh harmonic, which produces a rough and discordant tone, is carefully eliminated by piano makers. It is these harmonics that produce the characteristic tone of the piano and differentiate it from a harp, for example. When broadcast listeners hear a piano that sounds like a harp they may know of a certainty that some of the component frequencies of the tone are excluded or distorted.

Each instrument in an orchestra has a characteristic timbre or tone quality depending on the number and relative strength of the harmonic overtones. The tone of a flute impresses the listener as being pure and limpid because of its extreme simplicity. It has only the fundamental and the octave or second harmonic. The French horn is next in purity, including the third harmonic. On the other hand the tones of the violin, oboe, clarinet, saxophone, etc., are very rich in overtones. To bring out the characteristic timbre of these various instruments the harmonic overtones are absolutely essential.

In the case of percussion instruments such as drum, cymbals, xylophone, castanets etc., the tones are even more complex, for here we have to deal, not with steady harmonic overtones, but with transients of an abrupt and impulsive character. These transients, like others with which the electrical engineer is familiar, involve a whole series of component overtones, and the sharper and more abrupt the pulses the higher the frequencies required to represent them faithfully. If the high frequency overtones are suppressed the transients are rounded and blurred and they lose their sharp brilliancy. The clean cut reproduction of a jazz band, whether wholly desirable or not, is a severe test of the fidelity of a radio receiver.

It is thus clear that the range of frequencies that faithful radio phone reception must include is far greater than that indicated by the fundamental frequencies of the notes in the musical scale.

In the case of articulate speech, overtones, both harmonic and inharmonic, are even more important. The vowel tones are relatively simple, including, like the tones of musical instruments, various harmonic overtones which determine the vowel. Consonants are very much more complex and involve high overtones. The sibillant consonants such as *s*, *z*, *ch* and *zh* involve extremely high sustained frequencies. Other consonants such as *p*, *b*, *t*, *d* involve abrupt pulses which have essentially the nature of transients and like other transients, when analyzed into their components, involve very high frequencies. When the high frequencies are excluded these transients lose their characteristic abruptness and become rounded or blurred and so are unintelligible. Everyone is familiar with the frequent necessity of spelling "C— for Charles, A— for Arthur, T— for Tommy" when talking over the telephone. If the high-frequency overtones were all transmitted this would be unnecessary.

Just how much of the total range of overtones is required for acceptable reproduction is a question of compromise and individual judgment. All the overtones are desirable, but perhaps not all are necessary. The average human ear hears frequencies as high as 15,000 cycles, though extremely high frequencies are usually unimportant. Frequencies as high as 10,000 cycles are of material value, and it is generally considered that frequencies as high as 6,000 or 8,000 cycles are necessary in the faithful reproduction of tone and speech. To avoid needless debate it would seem proper to place the desirable upper frequency somewhere between the limits of 6,000 and 10,000 cycles per second.

The lower limit of frequency is unimportant here, since it concerns chiefly the design of the audio-frequency amplifier and sound projector, which are outside the scope of the present paper. This fact is pertinent, however. With the advent of improved audio-frequency amplifiers and loudspeakers the lower end of the scale has been greatly extended, but when a good audio system is attached to an ordinary radio receiver the results are often disappointing. There is a preponderance of bass and the tone has an unreal, hollow sound—the "rain-barrel" effect. It is not safe to blame this on the audio system. Usually it is due simply to the lack of sufficient overtones to balance the bass. When the overtones are supplied by the radio receiver the tone becomes normal, with its full richness and with a brilliancy that must be heard to be appreciated.

The Width of the Modulation Bands and their Relation to Selectivity

Now when a given band of audio frequencies is used to modulate a high-frequency carrier wave, a corresponding series of frequencies appears in the modulated wave. The various frequencies occur in the same arithmetical progression but stepped up to a higher level, *i.e.*, for each frequency f of the modulation there is a corresponding radio frequency $F+f$, F being the carrier frequency. These summation frequencies make up the upper side band. There is also an inverted arithmetical series of frequencies equal to $F-f$, which constitutes the lower side band. The width of each side band, measured in kilocycles, is equal to the highest frequency included in the modulation. Thus if we assume that the modulation includes frequencies up to 10,000 cycles per second, the total width of the transmitted spectrum, including both side bands, will be 20 kilocycles. If the highest frequency of the modulation is 5,000 cycles the total width of the transmitted spectrum is 10 kilocycles.

The two side bands are not both necessary for distortionless transmission. Either may be suppressed without affecting the fidelity of reproduction, and in some transmitting systems one side band is suppressed; but when both are transmitted, each must be received in its entirety. If either side band is trimmed, the symmetry of the reception is destroyed and distortion results.

The condition of distortionless reception and faithful reproduction of tone is that all the frequencies in the transmitted band of the modulated wave shall be received in their true relative intensities.

On the other hand *the condition of selectivity requires the effective separation of all the frequencies of an interfering wave from those of the desired signal.*

Obviously, if the side bands of two interfering waves overlap the two conditions of fidelity and selectivity are incompatible. If they are closely adjacent but not overlapping the problem resolves itself into securing uniform reception of all frequencies within the modulation band, with a sharp cut-off at the extremities of the band.

The legalized broadcast channels at the present day include carrier frequencies differing by as little as 10 kilocycles, but the geographic locations are so assigned that such closely adjacent frequencies are widely separated so that the problem of overlapping

bands usually is not serious, but interference between waves differing by 20 kilocycles is not unusual.

If we are to receive a full frequency band or spectrum of 20 kilocycles width, including modulation frequencies up to 10,000 cycles, obviously the bands of two modulated waves 20 kilocycles apart will meet with no space between. To receive without distortion one wave band of 20 kilocycles width and exclude the

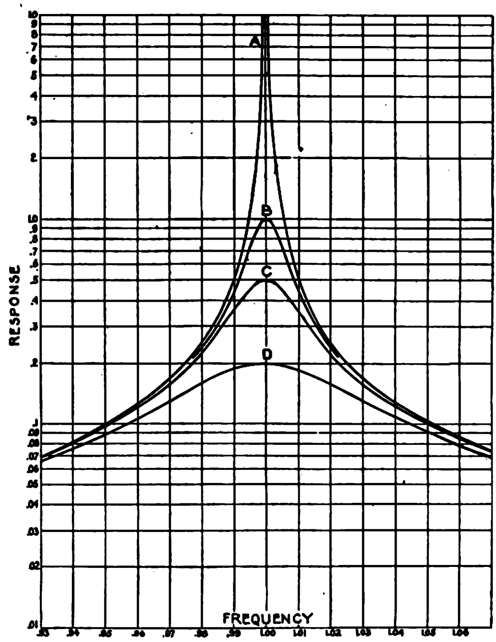


Fig. 1—Generalized Frequency Characteristic (computed) for a Single Resonant Circuit.
Graph A—for an ideal circuit with zero damping.
Graph B—for a circuit having an effective resistance equal to 1 per cent of the reactance at resonant frequency.
Graph C—for 2 per cent effective resistance.
Graph D—for 5 per cent effective resistance.

adjacent wave requires a receiver of extraordinary qualities, receiving with equal efficiency all the frequencies within the 20-kilocycle band and abruptly cutting off all frequencies outside this band. In other words it requires a receiver whose frequency characteristic of reception is substantially rectangular.

The receivers in general use today do not work that way. In the allied art of wire telephony filters of the Campbell type are

used to separate adjacent frequency bands. Similar means may be used at radio frequencies where the bands are fixed and where such complicated circuit networks are permissible, but *in the field of broadcast reception there is one added condition that must be met, namely the feasibility of adjustment of the band of reception in the frequency scale.* This requires simplicity of the circuits and apparatus and the operation of all frequency adjustments by a single control.

To meet these three conditions of fidelity, selectivity, and simplicity was the object of the research set forth in this paper.

Limitations of Selectivity by Resonance

Actual broadcast reception, as practised today by the ordinary methods, falls far short of meeting these conditions. A receiver whose selectivity depends on the use of tuned resonant circuits operated in synchronism, such as the tuned radio-frequency amplifier, has an overall frequency characteristic determined by the well-known resonance curve.

The ideal resonance curve of a single circuit without damping is shown at *A* in Fig. 1. The curve of such an ideal circuit rises to infinity at the resonance point, and at frequencies above and below resonance follows a hyperbolic law. When resistance or other damping is introduced into the circuit the current is always finite. At the resonance point, where the total reactance is zero, the current is determined by Ohm's law, assuming all the energy losses in the circuit to be included in its effective resistance. The peak of the resonance curve thus comes down to a point *B*, Fig. 1. At frequencies above and below resonance the curve drops off less abruptly than in the case of zero resistance and the actual curve approaches the ideal as the departure from resonance frequency increases in either direction. The graphs *B*, *C*, *D*, of Fig. 1 are a family of resonance curves plotted for effective resistances equal to 1 per cent, 2 per cent and 5 per cent respectively of the capacity reactance or the inductance reactance at resonant frequency. These figures represent that may properly be called the power factor of the resonant circuit. In Fig. 2 the same thing is shown by oscillograph records of the current in an actual resonant circuit.

This family of curves illustrates the effect of damping on selectivity. Damping causes a large diminution in the peak value

of the resonance curve and a smaller diminution at points off resonance. *It will be noted that damping does not broaden the resonance curve. It merely flattens it.* This fact is sometimes overlooked.

When a resonant circuit is used as a selective element in a radio receiver, the amount of distortion is measured by the dropping off of the frequency characteristic curve, Fig. 1, within the range of frequencies included in the modulation band. The ideal circuit with no damping would not receive any modulation at all, since the ratio of the current strength at carrier frequency to that at any of the side band frequencies is infinite. This explains the poor tone quality of a regenerative receiver, since the effective

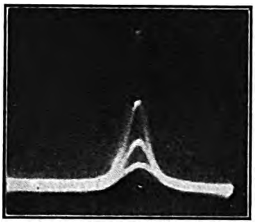


Fig. 2—An Oscillograph Record Showing a Family of Response Curves Corresponding to Fig. 1, for an Actual Resonant Circuit.
Graph A—with no external resistance, effective resistance of the circuit being 0.5 per cent of the reactance at resonant frequency.
Graph B—the same circuit with an added ohmic resistance of ½ per cent.
Graph C—the same circuit with an added ohmic resistance of 1 per cent.
Graph D—the same circuit with an added ohmic resistance of 2 per cent.

resistance is reduced to a very small value by regeneration and the response curve approaches the ideal.

Fig. 3 is a similar family of graphs showing the characteristics of two resonant circuits in cascade and tuned to synchronism. Because of the geometric property of such circuits the ordinates of each graph for a two-circuit system are equal to the squares of the corresponding ordinates for a one-circuit system. The graphs are plotted on a logarithmic scale of ordinates so that the graphs of Fig. 3 are derived from Fig. 1 by simply doubling all ordinates, the ordinates being measured from the resonance point, so that the peak values of all curves coincide, the response of the system at resonance being taken as unity. The value of the

ordinate at any frequency will then measure the response of the system at that frequency, relative to its response at resonance.

Fig. 3 shows a similar family of graphs for three synchronously tuned circuits in cascade. These are obtained from Fig. 1 by trip-

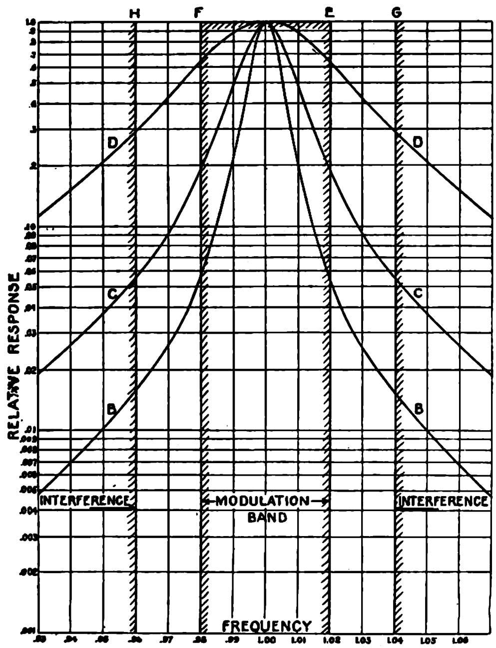


Fig. 3—Response Curves of a System of Two Resonant Circuits in Cascade, Synchronously Tuned. Derived from Fig. 1 by Squaring the Ordinates and Reducing to a Common Origin.

Graph B—for effective resistance = 1 per cent.
Graph C—for effective resistance = 2 per cent.
Graph D—for effective resistance = 5 per cent.

The lines *E*, *F* represent the boundaries of the side bands of a modulated wave with a limiting modulation frequency of 10,000 cycles and a carrier frequency of 500 kilocycles.

The lines *G*, *H* represent the frequency of an interfering wave differing by 20 kilocycles or 4 per cent from the carrier frequency of the signal wave.

The intersections of these lines with the several graphs give the relative reception of the various frequencies.

ling all ordinates, and reducing the curves to a common origin, as in Fig. 3.

These graphs are perfectly general and independent of any particular values of inductance, capacitance, or frequency, and they do not involve any assumptions as to whether the circuits

are coupled by amplifying tubes or otherwise, or as to the degree of amplification. The frequency coordinates are plotted on a linear horizontal scale, the frequency at resonance being taken as unity. The vertical coordinates, which represent the relative response of the system to a given impressed electromotive force at the different frequencies, are plotted on a logarithmic scale, the unit of the

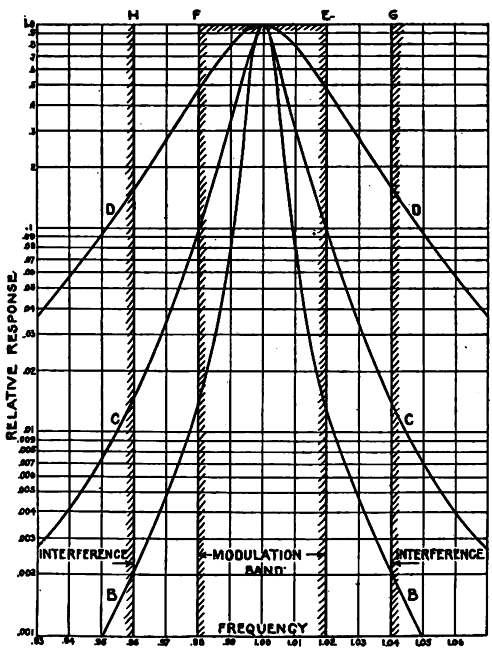


Fig. 4—Response Curves of a System of Three Resonant Circuits in Cascade Synchronously Tuned. These Circuits Are Derived from Fig. 1 by Cubing the Ordinates and Reducing to a Common Origin.
Graph B—for effective resistance = 1 per cent.
Graph C—for effective resistance = 2 per cent.
Graph D—for effective resistance = 5 per cent.
The dotted lines *E, F, G, H* correspond to the similar lines of Fig. 3.

scale being the current set up in the last circuit of the system at resonant frequency. They can be applied to any particular case by substituting the proper numerical values.

Assume for example that it is desired to find the amount of distortion in a two-circuit receiver with 1 per cent power factor when receiving a signal band whose carrier frequency is 500 kilocycles and whose side bands extend 10 kilocycles on each side of the

carrier. The frequency at the upper limit of the band is then $\frac{510}{500}=1.02$, and at the lower limit is $\frac{490}{500}=0.98$. The vertical dotted lines E, F drawn at these points on the frequency scale of Fig. 3 thus represent the band width of the modulated signal wave. The intersection of these lines with the graph B for 1 per cent gives the relative current strength at the extreme side-band frequencies. This is found to be 0.055, the current at carrier frequency being 1.0. The received signal strength at the extreme side-band frequency is thus only $\frac{0.055}{1.0}$ or 5.5 per cent of that which would be required for distortionless reception. This amounts practically to the elimination of the extreme side bands. The received signal strength for a

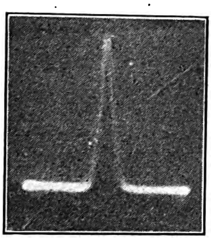


Fig. 5—Oscillogram of the Frequency Characteristic of a Typical Broadcast Receiver, Showing its Sharply Peaked Form, Corresponding to a Typical Resonance Curve, Fig. 4.

The horizontal coordinate is the frequency scale, the vertical coordinate is the scale of response to unit impressed electromotive force at the various frequencies.

This graph, like Fig. 2, differs from the computed curves in having a linear scale of ordinates while the computed curves have a logarithmic scale.

modulation frequency of 5,000 kilocycles is 0.21 or 21 per cent, hence the distortion even at moderate frequenciesis serious.

Again, suppose the condition of selectivity requires excluding a frequency differing from the signal carrier by 20 kilocycles or 4 per cent. Drawing another pair of lines G, H at frequencies 1.04 and 0.96 respectively, the intersection of these lines with the 1 per cent graph B shows that the interfering signal strength will be 0.015 or 1.5 per cent of the signal carrier frequency. It thus appears that while the selectivity may be considered passable, the side-band frequencies are reduced to the point of serious distortion.

If we attempt to remedy this defect by the introduction of damping we find that with 5 per cent damping the side-band recep-

tion is 0.62 or 62 per cent of the carrier, which might be acceptable, but the receptional interfering frequency has risen to 0.29 or 29 per cent of the reception at carrier frequency, which is no selectivity at all. At the same time, the sensitivity at carrier frequency has fallen from 1.0 to 0.04, or only 4 per cent of the sensitivity with 1 per cent damping.

Taking any intermediate value of damping these two factors will differ in relatively lesser or greater degrees, but in no case can a reasonable approximation to full band reception be secured without an undue sacrifice of selectivity.

Making similar assumptions for a three-circuit system, the similar lines *E*, *F*, *G*, *H* of Fig. 4 show that, for 1 per cent power factor, the 10-kilocycle side band has a relative intensity of 1.3 per cent while the 20-kilocycle interfering frequency is 0.2 per cent.

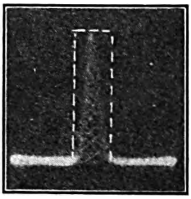


Fig. 6—The Same Graph Shown in Fig. 5 Superimposed on a Rectangle Representing the Modulation Band of a Wave Having a Limiting Modulation Frequency of 10,000 Cycles.

The shaded area illustrates the effective reception of the set, the unshaded area represents the portion of the side bands that is eliminated.

Even for a 5-kilocycle side band the relative intensity is only 9 per cent, showing a degree of distortion that is prohibitive.

The graph for 5 per cent damping shows that the 10-kilocycle side band has a relative intensity of 46 per cent, as compared with 62 per cent, for the two-circuit system, while the interfering signal strength is decreased from 29 per cent to 15 per cent, which is still far too large for satisfactory seletivity.

A study of these graphs shows clearly that any attempt to improve fidelity of reception by the introduction of damping does not change the essential form of the frequency characteristic, but merely flattens it. The characteristic is always essentially curved and the result at best is merely to mitigate the evil. It is impossible to obtain distortionless reception in this way. Furthermore any improvement in the fidelity of reception is accompanied by a cor-

responding loss of selectivity, and reasonable selectivity cannot be secured without a considerable distortion. The essential limitation of a system including synchronously tuned circuits is thus evident.

As a practical illustration of the way this works out, Fig. 5 shows by an oscillogram the frequency characteristic of a receiver of well-known and very popular make. The width of this curve at its base is of the order of 15 kilocycles, the carrier frequency being 600 kilocycles, indicating good selectivity, but the peak of the curve is so sharp that the side bands are severely trimmed. The resulting distortion of tone is clearly evident in the output of the receiver.

In Fig. 6 this curve is shown superimposed on a dotted rectangle representing the modulation band of a signal wave having a 10-kilocycle limit. It is readily seen how small a fraction of the side bands is effective.

It is thus clear that any attempt to improve the quality of reception by flattening the resonance curve by means of damping, results at best in mitigating the distortion but can never entirely remove it, and it is accompanied by a serious loss in selectivity.

If full-band reception is to be secured with reasonable selectivity, clearly something more than the usual multiple resonance methods is required.

The Possibilities of Band Reception

The ideal solution of the problem is a system which has a substantially rectangular frequency characteristic, that is, one which gives substantially uniform reception over a definite band of frequencies, including all the side-band frequencies of the modulated wave, with a sharp cut-off for frequencies outside this band. Such a system in its ideal form would give distortionless reception, with a selectivity that is limited only by the overlapping of an interfering wave with the signal band. The closeness with which this ideal has been approximated will appear later.

The problem admits of two distinct solutions.

1—The use of one or more band selectors, each of which possesses a substantially rectangular frequency characteristic.

2—The use of two or more receiving elements whose individual frequency characteristics are not rectangular but which in combination have an overall characteristic that is substantially rectangular.

Again various combinations of these elements are possible. For example we may have—

a—A band selector with a flat amplifier.

b—A band amplifier.

c—Various combinations of band selector and band amplifier.

All of these solutions have been developed into practical working receivers.

The Band Selector

A band selector has been developed which meets the three stated conditions *i.e.*, fidelity, selectivity, and simplicity. It comprises in general a system of reactances so related to each other that they are mutually balanced, not merely at a single

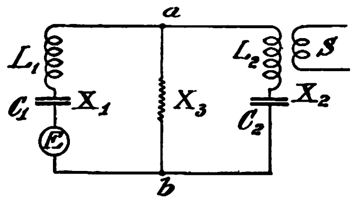


Fig. 7—Generalized Schematic Diagram of a Band Selector.
X_1 and X_2 are two reactive couples. X_3 is the bridging or band-forming reactance. E represents the impressed signal electromotive force. S represents the output.

frequency as in the case of the ordinary tuned circuit, but also at any frequency within a given band. At any frequency outside of this band the reactances are not balanced and the unbalanced reactance is high. As a result of this property, the band selector unit responds with substantial equality to all frequencies within its characteristic band and is non-responsive to frequencies outside this band. When the system is suitably designed the cut-off at the limits of the band is very sharp. The electrical and mechanical construction is exceedingly simple, and frequency adjustment is obtained by means of only two variable elements operated by a single control.

The band selector is shown in generalized form in Fig. 7. It employs two reactive couples X_1 and X_2, preferably alike, each having a capacitance and an inductance that are balanced within themselves at the same frequency, together with a third reactance X_3 which is common to both. This third reactance is small in relation to the reactances of the two reactive couples and may be either inductive or capacitive.

An input electromotive force is impressed on the system in any suitable way as at E and the output is taken off in any suitable way as at S. At a particular frequency F_1, this being the frequency at which the reactances of the couples X_1 and X_2 are balanced within themselves, the overall reactance of the circuit including X_1 and X_2 will be zero, current at the frequency F_1 will circulate through the branches X_1 and X_2 without traversing X_3, and the system has zero reactance at this frequency.

At any other frequency the reactive couples X_1 and X_2 will not be balanced within themselves. The result will be a potential difference across points a, b, the terminals of the bridging reactance X_3. If the frequency is lower than F_1 the reactancs of X_1 and X_2 will be capacitive. If now the reactance X_3 is inductive, it will tend to neutralize the unbalanced capacitance of branches X_1 and

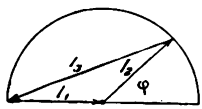


Fig. 8—Vector Diagram Showing the Phase Relations of the Currents in the Band Selector.

I_1, I_2, and I_3 represent the currents in the Branches X_1, X_2, and X_3 respectively. φ is the phase difference between I_1 and I_2, which varies from zero to 180 deg. The semicircle is the locus of the apex of the triangle formed by I_2 and I_3, I_1 being fixed.

X_2, provided their combined reactance is no greater than X_3. In that case current will flow through X_3 of such amount that the reactive electromotive force across points a, b, due to the current in X_3, is equal to that due to the currents in X_1 and X_2. The phases of the currents in X_1 and X_2 will adjust themselves so that I_3 is equal to the vector sum of I_1 and I_2.

The operation of the band selector unit may be more readily understood by reference to the vector diagram Fig. 8. Let the currents set up by the impressed electromotive force E in the three branches X_1, X_2, and X_3 be I_1, I_2 and I_3 respectively. These three currents are considered positive when they flow in the direction from the common point a of the branches to the common point b. Since the total current flowing into or out of points a and b must be zero, the current I_3 in the common reactance X_3 must be equal and opposite to the vector sum of currents I_1 and I_2 in the other two branches. This relation is shown by the vector diagram Fig. 8, I_1 being regarded as fixed and I_2 rotating with relation to I_1, with

the phase difference ϕ. For any value of ϕ the current indicated by the vector I_3 is the third side of the triangle formed by I_1 and I_2.

In the case of a symmetrical system where the branches X_1 and X_2 are alike, the currents I_1 and I_2 will be equal and the current I_3 will have the value:

$$I_3 = -I_1 - I_2 = -2I_1 \cos \frac{\phi}{2}$$

It thus appears that the current in the common reactance I_3 varies between the limiting values $-2I_1$ and zero, as the phase difference ϕ between currents I_1 and I_2 varies from zero to 180 deg.

This phase relation depends upon the frequency of the impressed electromotive force in the following manner. Since the points a, b, are points of like potential difference in the three branches of the system, the current distribution in the system must be such that the reactive electromotive forces in the three branches are equal, that is

$$x_1 I_1 = x_2 I_2 = x_3 I_3$$

In the case of symmetry this becomes

$$x_1 I_1 = x_2 I_2 = -2x_3 I_1 \cos \frac{\phi}{2}$$

or

$$x_1 = x_2 = -2x_3 \cos \frac{\phi}{2}$$

For the limiting case where ϕ equals zero this becomes

$$x_1 = x_2 = -2x_3$$

For the other limit where ϕ equals π or 180 deg.,

$$x_1 = x_2 = o.$$

In other words there is a limiting frequency F_2 at which the reactances are balanced when the phase difference between I_1 and I_2 is zero and the current in the common reactances X_3 becomes equal to twice I_1, and there is another limiting frequency F_1 at which the reactances are balanced when the currents I_1 and I_2 differ in phase by 180 deg. and the current in X_3 becomes zero. This latter frequency is the one at which the reactances of the reactive couples X_1 and X_2 are balanced in themselves, and the current

circulates wholly through the branches X_1 and X_2, no part of it traversing X_3. At the former frequency the reactances of the reactive couples X_1 and X_2 are not balanced in themselves, but they are completely balanced by the reactance X_3 when the entire current of both branches flows through X_3.

At any frequency between these limits, the phase difference between the currents I_1 and I_2 will have a value lying between zero and 180 deg., the resulting current I_3 will have a value intermediate between $-2\ I_1$ and zero, and the reactive electromotive force across X_3 will have a corresponding intermediate value equal to the reactive electromotive force in the branches X_1 and X_2.

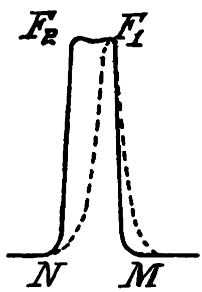


Fig. 9—Frequency Characteristic (computed) of the Band Selector Shown in Fig. 7.

F_1 is the first limiting frequency, where $\varphi = 180$ deg., and $I_3 = 0$. F_2 is the second limiting frequency where $\varphi = 0$, and $I_3 = 2I_1$. The dotted curve is the characteristic of the same system with X_3, open circuited. It is a typical damped resonance curve whose peak frequency is equal to F_1.

Note the steep gradient of the cut-off, F_1M and F_2N, and the sharpness of the bend at the base compared with the resonance curve.

The reactances of the system as a whole are thus completely balanced at any frequency between the limiting values F_1 and F_2, and the system will transmit freely any frequency in a band comprised between these limits. When X_3 is an inductance the frequency F_2 is lower than F_1. When X_3 is a capacitance F_2 is higher than F_1.

Since the phase difference between I_1 and I_2 cannot be less than zero or greater than 180 deg., if the impressed electromotive force has a frequency lower than the limiting frequency F_2 or higher than the limiting frequency F_1; (or vice versa when X_3 is capacitive) there is no possible phase adjustment which will cause the reactances to balance, hence there will be an unbalanced reactance in the system that will prevent the flow of current.

The above analysis neglects the resistance of the system. If the resistance and other losses are low, as they should be, the cut-off at the limiting frequencies is very sharp, and the frequency characteristic of the band-selector unit has the form shown in Fig. 9.

The width of the band depends upon the relation of the reactance x_3 to the other reactances of the system. Thus if X_3 is an inductance, the band width depends upon the relation of this inductance to the inductances L_1 and L_2. If the reactance X_3

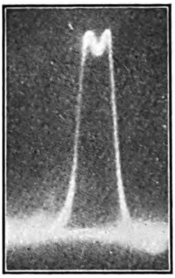


Fig. 10

Fig. 11

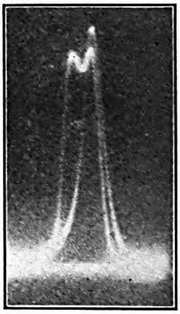


Fig. 12

Fig.10—Oscillogram Showing the Frequency Characteristic of an Actual Band Selector Corresponding to Fig. 9. Carrier Frequency, 635 kc. Effective Band Width, 18 kc.

Fig. 11—Oscillogram Showing the Frequency Characteristic of the Same Band Selector Converted into a Simple Resonant Circuit by Opening the Reactance X_3.

Fig. 12—Composite Oscillogram, the Graphs of Figs. 10 and 11 Being Super-imposed on the Same Film.

Note that the width of the band characteristic is equal to the width of the resonance characteristic at a point above the base. As the zero axis is approached the band characteristic becomes much narrower than the resonance curve, showing improved selectivity.

is a capitance the band width is determined by the relation of the capacity reactance of X_3 to the capacity reactance of C_1 or C_2.

It is of interest to note the relation of the frequency characteristic of the band-selector unit to the characteristic of a tuned resonant circuit. Thus if the common or bridging reactance X_3 is omitted the two branches X_1 and X_2 together constitute a resonant circuit tuned to a certain frequency F_1; this being one of the limiting frequencies of the band of the selector unit. The resonance curve of such a tuned circuit is shown by the dotted lines in Fig. 9 in its characteristic sharply peaked form.

When the common reactance X_3 is added to the system the curve takes the band form shown in full lines, the limiting frequency F_1 corresponding to the natural frequency of the tuned circuit and the limiting frequency F_2 being below or above this frequency, depending upon whether the reactance X_3 is inductive or capacitive.

When the reactance X_3 has a suitable small value in reference to the other reactances the widths of the two curves at the base are substantially the same, showing that the uniform band reception is achieved without any loss in selectivity, but rather with a noteworthy gain, as will now appear.

The frequency characteristic of an actual selector of this type is shown in the oscillograph record Fig. 10. It will be noted that the band is substantially rectangular, the sides being almost vertical. The gradient of the cut-off is very much sharper than that of a resonant circuit made up of similar reactances. For comparison Fig. 11 shows a true resonance curve obtained with the bridging reactance X_3 removed, in which case the system X_1, X_2 becomes a simple resonant circuit. In Fig. 12 the two graphs are superimposed on the same film.

When the band selector is completed by inserting the reactance X_3, the point F_1 remains fixed and the cut-off from F_1 to M becomes very much steeper, corresponding to the cut-off from point F_2 to N. The width of the curve at its base is substantially equal to that of the resonance curve notwithstanding the great width of the band at its top.

The full gain in selectivity is not clearly seen from the films, but it will be noted that the cut-off lines drop straight to a point close to the zero axis, into which they merge by a sharp bend. In the resonance curve Fig. 11 the approach to the zero axis is gradually rounded. This bend of the characteristic is the factor that chiefly determines selectivity. A sharp bend of the curve at these points means a small value of an interfering current.

The superior selectivity of the band selector is thus evident. In fact, a single band selector has a selectivity about equal to two resonant circuits made up of the same coils and condensers.

It will be noted that the amplitude of the transmission in the band selector is substantially the same as that of a resonant circuit having the same elements, notwithstanding the greatly widened band. In other words, the band selector broadens the

scope of the reception without any loss in signal strength. This is in marked contrast with the results obtained by damping a tuned circuit in an effort to improve the fidelity of reception. In that case any gain is accompanied by a flattening of the response curve, i.e., reducing its amplitude, and at the same time reducing its selectivity, as shown in Fig. 2, and the result is, at best a makeshift.

With the band selector the characteristic is broadened, giving perfect (not approximate) side-band reception, with no loss in sensitivity and with a great gain in selectivity.

Applications of the Band Selector

The generalized band selector Fig. 7 may be readily adapted to radio reception by antenna or loop. Such an arrangement

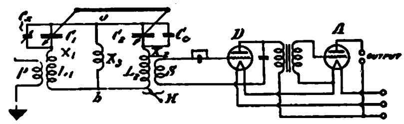


Fig. 13—Schematic Diagram of a Broadcast Receiver Including a Single Band Selector X_1, X_2, X_3 Coupled to an Antenna A and Feeding an Aperiodic Amplifying System, D, A.

Cc is the fixed antenna compensator, *Cx* is a variable condenser for adjusting the compensation.

using an antenna is shown in Fig. 13. The antenna coil P, is coupled to the band selector by the coupling P, L, a, preferably with a step-up ratio. The capacity introduced into the branch X_1, by the antenna is compensated by a fixed capacity C_c in the branch X_2. In order to permit compensation for an antenna of any desired capacity without disturbing the frequency calibration of the system an additional variable condenser C_x is inserted, which makes up the difference between the capacity introduced by the antenna and the capacity C_c, so that the symmetry of the system is secured.

The adjustment of C_x is made arbitrarily until the signal strength becomes maximum, after which no further adjustment of C_x is required for a given antenna. The sole frequency adjustment is that of the two coupled condensers C_1 and C_2.

It is of interest to note that any lack of symmetry in the system that might result from careless or imperfect adjustment or the capacity C_x, within reasonable limits, does not materially alter the band form of the characteristic but merely reduces its amplitude.

The band selector may be used in a variety of ways. It may be employed as the sole selective element of a receiving system, feeding a flat amplifier, as shown in Fig. 13. This makes a system of great simplicity and high efficiency, and with sufficient sensitivity and selectivity for ordinary broadcast reception. It is particularly adapted to use in the metropolitan areas.

The band selector lends itself readily to use as an interstage coupling element of a radio-frequency amplifier. Such an arrangement is shown in Fig. 14.

In this arrangement each of the selective elements has in itself a substantially rectangular band characteristic. With two or more such units combined in an amplifier the over-all frequency characteristic has a similar rectangular form with a sharper cut-off.

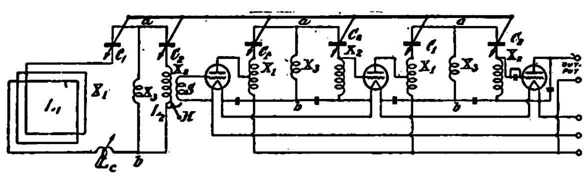


Fig. 14—Schematic Diagram of a Receiver Including Three Band Selectors. The first selector includes a collecting loop L_1 and the second and third selectors are used as interstage couplings in a radio-frequency amplifier.

The inductances X_1 and X_2, and the capacitances C_1 and C_2 are all alike, hence frequency selection is accomplished by a single control, coupling the equal condensers C_1 and C_2.

Lc is a small compensating inductance to adjust any inequality between the inductance of the loop L_1 and that of the selector coil X_2.

Increased sensitivity and selectivity are thus secured with no diminution in the band width. This is in marked contrast with the tuned radio-frequency amplifier, where an increase in the number of stages inevitably narrows the characteristic. When the various band selector units are made alike, as they may readily be, the whole system, including the compensated antenna selector, is symmetrical, and all the variable elements may be operated by a single control, as shown.

The Spaced Band Amplifier

In the band amplifier of the second type the several stages have different frequency characteristics which are not rectangular in themselves, but in combination they produce an over-all band characteristic. This is done in the manner illustrated in Fig. 15 where 1, 2, 3 are of the individual characteristics of a three-stage

amplifier. These will have in general the characteristic form of a damped resonance curve, and they are made similar but differently spaced in the frequency scale. When the three stages are combined in an amplifier the over-all characteristic does not follow the geometric law, as in the case of synchronously tuned stages, but has a form totally different from that of the resonance curve. In general there is a certain spacing at which the over-all amplification becomes substantially constant over a considerable frequency band, and drops off abruptly with a sharp cut-off at the extremities of the band, as shown in curve 4 Fig. 15. The cut-off is very much

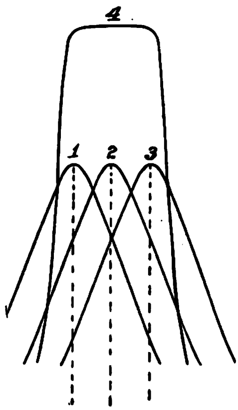


Fig. 15—Graphs Illustrating the Principle of the Spaced Band Amplifier. Graphs 1, 2, 3 represent the frequency characteristics of three successive amplifier stages. These characteristics are similar except that they are spaced in the frequency scale.

Graph 4 shows the overall characteristic of an amplifier whose individual characteristics are 1, 2, 3.

It will be seen that the overall characteristic is substantially flat over a definite band, with a sharp cut-off at the extremities of the band.

sharper than that of the component characteristics; in fact it is much sharper than the gradient for a three-stage synchronously tuned system of the same damping. When the circuits are made with small damping and the correct spacing the characteristic is substantially rectangular. If the spacing is closer than the optimum value the over-all characteristic will have a bump in the middle. If the spacing is greater than the optimum there will be a double peak. The value of the optimum spacing depends upon the form of the individual characteristics.

The spacing may be secured in a variety of waves. Thus the coupling coils or transformers may be made with different inductances and the frequency adjusting capacities made alike, in which case the band width, measured as a fraction of the carrier frequency, will be uniform over the range of the frequency adjustment. Or the transformers or inductive elements may be made alike and the capacities different, or both inductances and capacities may be made alike and a small spacing inductance added to each of the lower frequency stages. From a practical standpoint it is usually desirable to make the coils and capacities alike and to add spacing inductances or spacing capacities to the lower frequency stages, as shown in Fig. 18.

The graph in Fig. 15 is a typical over-all frequency characteristic of a three-stage amplifier, this curve being derived by

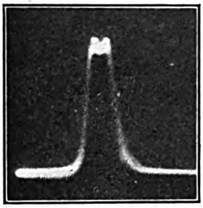


Fig. 16—Oscillogram Showing the Frequency Characteristic of an Actual Spaced Band Amplifier of the Type Shown in Fig. 15.
Note the steepness of the cut-off and the sharpness of the bend at the zero axis.

computation. A photographic oscillogram made from an actual amplifier of this type shown in Fig. 16.

The band characteristic Fig. 16 may be compared with the peaked characteristic of a standard commercial receiver of the tuned radio-frequency type shown in Fig. 5.

It will be noted that, although the band amplifier curve of Fig. 16 is substantially flat for a band width of 20 kilocycles (the carrier frequency in this case being 600 kilocycles,) the cut-off is so much sharper than that of the tuned radio characteristic Fig. 5 that the widths of the curves at the base are substantially the same, but the bend is sharper in the case of the band amplifier. It is this width and bend that determine the selectivity of the system. It thus appears that a band amplifier giving a band width of 20 kilocycles is fully equal or superior in selectivity to receivers of the geometrically tuned type.

The spaced band principle may be applied to a great variety of constructions, including the usual type of tuned radio-frequency amplifier. Fig. 17 shows an oscillogram obtained from the same receiver that gave the sharply peaked characteristic Fig. 5, by the simple expedient of suitably spacing the several stages. It will be noted that the width of the curve at the base is substantially

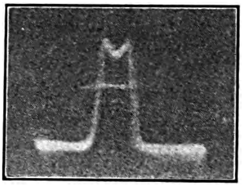


Fig. 17—Oscillogram Showing a Frequency Characteristic of the Same Amplifier Whose Normal Characteristic is Shown in Fig. 5, Now Converted Into a Band Amplifier by Spacing the Several Stages.

the same as in the case of the peaked characteristic Fig. 5 while the band width is sufficient to include a range of 15 kilocycles.

An important feature of the spaced band amplifier is its inherent stability. Since the several circuits are not synchronized, the tendency to regeneration and oscillation is small. In addition to its inherent stability other features are employed which make

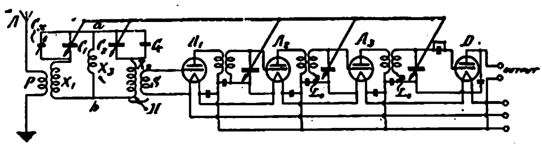


Fig. 18—Schematic Diagram of a Receiver Including a Single Band Selector Coupled to a Receiving Antenna and a Spaced Band Amplifier.
Ls, Ls are the spacing inductances.
The frequency adjusting capacities of the band selector and the amplifier are all coupled with a single frequency control.

the amplifier exceedingly stable. These features include an astatic winding of the coupling coils or transformers, which renders magnetic coupling between the stages negligible, and mutually reversed primary and secondary windings, which cause a phase reversal of any external electrostatic couplings, putting the potentials of the several stages in such phase relation that they do not cause regeneration. The resultant of these three features is an amplifier of such stability that no capacity neutralization or

balancing of any kind is required, and the coupling transformers may be placed closed together without shielding.

Various combinations of the band selector and band amplifier are possible. One of the most useful is shown in Fig. 18. The band selector is coupled to the antenna and feeds, through its output coil S, a spaced band amplifier of the type just described. In the arrangement here shown the three stages are all structurally similar and the spacing is accomplished by fixed spacing inductances L_s in two of the stages, as shown. Or two of the stages may have small spacing capacities.

In the design of the apparatus due attention is given to securing a suitable relation between the effective inductance and effective capacitance of the amplifier stages and the effective inductance and effective capacitance of the band selector elements. When this is done the frequency control curves of the various elements are readily synchronized so that the whole system may be operated by a single control means.

Practical Results

There has been much discussion as to the utility of the high overtones transmitted by the extreme limits of the side bands. The human ear is a long suffering organ, and will submit without protest to a degree of distortion of tone that is extraordinary. For this reason there is a great difference of opinion as to what degree of side-band reception is necessary to produce a passable result.

It is not the purpose of the present paper to enter into this discussion except to say that the development of the human ear to broadcast reception has been a progressive thing. At the beginning almost anything would be tolerated. As improvements have been introduced public taste has become more critical, and as each new improvement in tone quality has appeared the results that were formerly tolerated became unsatisfactory. Practical tests with listeners of all types and mental attitudes have shown without exception that when they are allowed to hear broadcast reception of a high quality, with all the overtones present, they are struck immediately by its superiority. The timbre of the various musical instruments takes on a brilliancy and a character that is startlingly realistic. The human voice is reproduced with all its subtilities, all the consonants being clearly articulated. The

"rain-barrel" effect that occurs when a good audio amplifier is combined with a selector that trims the side bands is entirely absent, and in its place is a balance of tone that is very satisfying to the much distressed feelings of a sensitive musical ear. With the correct tone balance, power amplifiers of any desired capacity may be used and the full round tone of an organ or an orchestra comes in with tremendous realism but with nothing to jar sensitive nerves.

It is believed that these results fully demonstrate the correctness of the theoretical considerations that have been set forth.

DIRECT COUPLED DETECTOR AND AMPLIFIERS WITH AUTOMATIC GRID BIAS*

By

Edward H. Loftin and S. Young White

Summary—A system for direct coupling of vacuum tubes to give composite detection and amplification is described. The system is designed to avoid frequency discrimination in amplifying audio frequency, and to be free from electrical and acoustical feed-back effects. High-μ tubes are used to control the output of the power amplifier through the aid of the very high filament-to-plate impedance of such tubes with extremely small expenditure of energy therein, and in such a way as to avoid effect of so-called "detector overloading." Microphonic effects are avoided. A unique method of automatically regulating the system to be responsive to carrier currents of different intensities is described, and includes an extension of the automatic effect to volume control. The entire system is designed to be inexpensive and extremely simple in the matter of construction.

It is the purpose of this paper to describe a composite detecting and amplifying system for modulated carrier currents having negligible frequency characteristics and minimum electrical and acoustical feed-back effects.

The fundamental system is shown in Fig. 1, where the effect of battery B_1 upon the grid of vacuum tube VT_2 is controlled by impedance of the plate-filament path of vacuum tube VT_1. Battery B_1 may be of the order of four volts. The resistance R_1 returns to positive filament. The potential which can develop

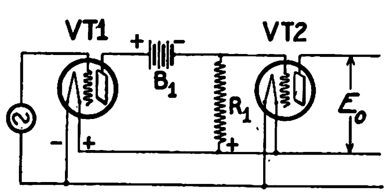


Fig. 1

across R_1 varies between zero and approximately $10\frac{1}{4}$ volts depending upon the potential of the grid of VT_1.

It will be noted that there is substantial freedom from coupling between the input and output circuits of each tube.

VT_1 operates at a very high plate-filament impedance value which results in small space current (order of 10 to 15 microamperes) with resultant long life of the tube and battery. When

* Original Manuscript Received by the Institute December 20, 1927.

the tube is operated at this small space current mechanical movements of the elements of the tube have no appreciable effect upon the magnitude of the current with the result that negligible microphonic effects and acoustical feed-back occur. It will be noted there are no resonant points or phase shifting means employed in the coupling. It is particularly desirable to use so-called high-μ tubes to obtain substantial amplification.

The circuit of Fig. 2 is merely that of Fig. 1 feeding into a power tube VT_3. Battery B_2 is selected to have the C-battery value required for the usual normal operation of the power tube plus about 50 per cent. Means to obtain a variable bias on the grid of the first tube are shown.

This system can be used as a combined rectifier and amplifier. A modulated carrier current impressed upon the input of the first

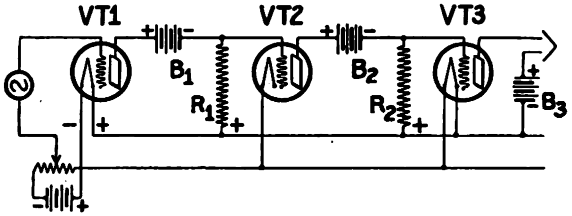


Fig. 2

tube lessens its plate-filament impedance, thus allowing B_1 to act more strongly on the grid of VT_2 as a negative bias to increase VT_2's plate-filament impedance and causing B_2 to be less effective on the grid of VT_3, all of which in turn causes the space current of VT_3 to rise in value. The current flow between VT_1 and VT_2 is mainly radio frequency, somewhat rectified, and this radio-frequency component persists even to the plate circuit of VT_3, this being demonstrated by the fact that tuned traps placed between any two tubes are quite effective in diminishing signal strength. Likewise any capacity to ground of the batteries lessens the sensitivity because of by-passing this persisting radio-frequency component.

The variable bias on VT_1 is necessary for several reasons. A strong carrier current, by reason of the rectifying properties of the system, will cause the plate current of the last tube to rise to saturation value, but this effect can be corrected by adding enough negative bias to VT_1. The bias is also needed to compensate for

the lack of uniformity of impedance of different high-μ tubes, these impedances having been found to vary widely in commercial tubes. This variable bias can also be used to compensate to some degree for variations in the *A* and *B* battery potentials on the tubes, and thus help to keep constant the steady component of the plate current of the last tube.

It is difficult to overload this system in the function of detection as the grid potential swing of the first tube is several times longer than the allowable grid potential swing of the second tube, which in turn is longer than the maximum allowable grid potential swing of the third tube.

The preceding system is principally of laboratory interest without some form of automatic control for the bias of VT_1. In a form

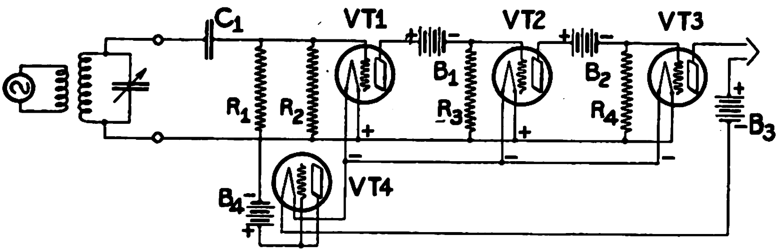


Fig. 3

of auto-compensator for the effect on the magnitude of the plate current of the last tube of varying intensities of carrier currents, it is possible to employ a rise of plate current above the normal to affect the grid bias of the first tube. However, the correction must apply only to the steady or direct current component of the plate current as affected by variations in tube impedances, carrier wave intensities, and the like, and not for plate current changes at an audio-frequency rate representing desired signal tones, as this would introduce audio-frequency feed-back, either positive or negative. One method suggesting itself is to take the voltage drop across a resistance in the plate circuit of the last tube and impress this drop on the grid circuit of the first tube. This requires filtering that introduces frequency discriminative. elements in the system, and does not hold the plate current in the last tube to the close limits desired. A more satisfactory auto-compensator is shown in Fig. 3.

VT_4 is the commercial tube type 199 so connected that its filament current is the total plate current drain of the radio-frequency and power tubes of a receiver, which current usually totals from 30 to 35 milliamperes. Anything tending to make the plate current of the last tube rise will allow this filament to warm up to the electron-emitting temperature to make the filament-plate path conductive, resulting in placing the potential of B_3 on the grid of VT_1. This filament-plate resistance of VT_4 is continuously variable from infinity to a comparatively small value with a change in filament current of a very few milliamperes, and consequently will place just enough negative bias on the grid of VT_1 to compensate for the strong carrier currents that may be encountered in receiving nearby or local stations.

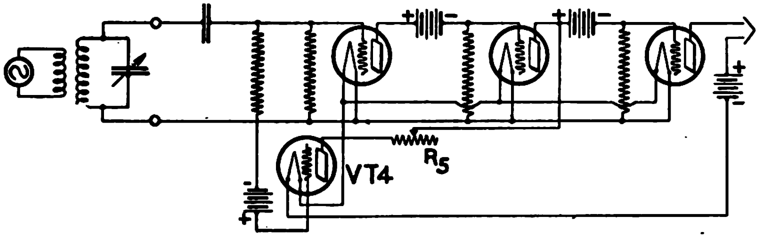


Fig. 4

With the tube so used the filament operates at very low temperature and consequent high thermal lag, which lag does not allow it to heat and cool rapidly enough to follow alternating currents even in the very low side of the audio range, and thus compensates only for effects lasting at least a substantial portion of a second.

R_1 must be smaller than R_2 for the reason that the voltage of B_3 is divided between VT_4, R_1 and R_2, and only the voltage across R_2 is effective in placing the bias on the grid. R_1 is useful only to lessen the effect of the capacity of B_3 to ground on the tuned circuit.

Fig. 5 shows a means for overcoming differences between current furnished by the high voltage plate supply and the current required to operate the 199-tube filament at its operating point in the neighborhood of 34 milliamperes. The resistance R_1 is a potentiometer of the order of one or two thousand ohms. When adjusted toward the minus side a local circuit is formed which

furnishes any additional steady heating current desired, to compensate for too little current from the plate supply. When adjusted toward the plus side its resistance is so low that its shunting action predominates, and equalization of too high a plate current is obtained.

The curve of Fig. 6 shows how bias due to B_4 becomes effective on the grid of VT_1 through variations in filament current of VT_4.

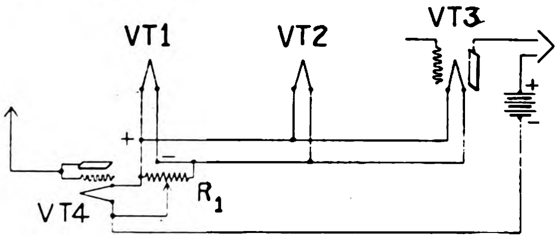


Fig. 5

This tube is usually operated at 34 milliamperes and will compensate for carrier wave intensities of 10 or 15 volts with a rise in plate current of VT_3, which will not exceed 3 milliamperes.

C_1 is used merely to isolate the input from any direct current potentials which might exist in the system to which the detector is

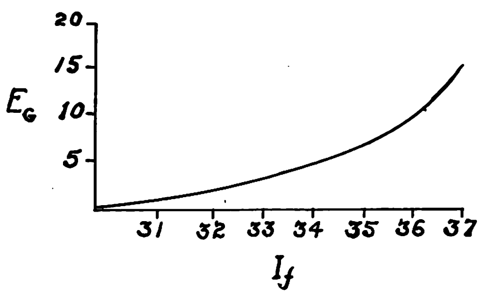


Fig. 6

connected. There is no grid rectification due to C_1 as the carrier wave causes an increase in conductivity of VT_1. If C_1 is small enough attempts at grid rectification will occur, which rectification will buck the plate rectification and weaken it. To prevent this C_1 must exceed 2000 $\mu\mu$f.

The 199-tube may also be used to prevent signal intensity from exceeding a predetermined limit in the manner shown in the

circuit of Fig. 4, where the grid of VT_4 is used for the correction of carrier current effect and the plate is used to introduce a comparatively low resistance in parallel with the plate circuit of VT_2. The impedance of VT_2, when handling a high intensity signal, is of the order of several megohms, so that placing a few hundred thousand ohms in parallel, as is done by the connection shown, results in a short circuiting effect that makes the variations less effective on the succeeding grid, thereby limiting volume.

The system may be used on alternating current throughout by isolating the filament sources to permit the use of a common high potential source for the *B* battery. Due to the unavailability of high-μ a-c. tubes this practice is not commercially expedient at the present time.

The sensitivity of the system is equivalent to the usual detector and two-stage transformer-coupled audio system.

The system described has proved very useful both in the laboratory and on the usual types of radio receiving sets. It is extremely simple and cheap to construct, and seems to avoid most of the troubles and disadvantages generally encountered in amplifying systems.

Discussion

Henry Shore: I think the answer to Mr. Dreher's question as to why the particular amplifier described is free from microphonic noises lies in the fact that the plate voltage is only four volts.

With such low voltage, the ionizing potential of the metallic plate is not reached and consequently, no secondary emission can take place. It happens that microphonics are, in the main, due to secondary emission and since this is absent with low-plate voltage, the amplifier is free from acoustical feedback.

ON ROUND-THE-WORLD SIGNALS*

By

E. O. Hulburt

(Naval Research Laboratory, Washington, D. C.)

IN his discussion of the measurement of E. Quäck of the time taken for the radio signal to go around the world[1] G. W. O. Howe has tacitly assumed, as Quäck did, that the ray goes roughly as shown in Fig. 1, that is, that the ray has the curvature necessary to keep it parallel to the surface of the earth. Using Quäck's time 0.137 sec. and Howe's group velocity calculation,

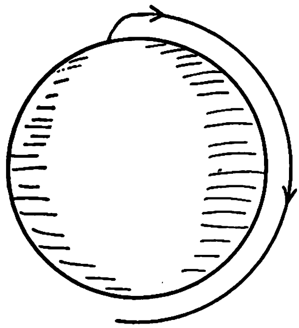


Fig. 1

we find that a 20-meter wave travels at a height of 48 miles above the earth, and the electron density at this height is 6.7×10^4; in the case of a 15-meter wave the height is 48 miles and the electron density 1.20×10^5. It seems possible that the round-the-world ray drawn in Fig. 1 (which is the ray of Howe's group velocity calculation), is a special case, and may not be the real path actually followed by the radio signal. Another hypothesis is to assume that the ray goes around in a sort of a polygon, as shown in Fig. 2. The sides of the polygon may or may not actually touch the earth at points *a*. Assuming that the sides of the polygon are close to tangency to the earth at points *a*, the length of the perimeter of the polygon is, from geometry

* Original Manuscript Received by the Institute, December 23, 1927.
* Published with the permission of the Navy Department.
[1] *Jahrbuch der drahtlosen Tel. und Tel.*, 30, 42, August, 1927.

$$\frac{2\pi r}{\theta} \tan \theta,$$

where r is the radius of the earth, 3970 miles $=6.38\times10^8$ cms. The perimeter is also equal to ct, where c is the velocity of light in vacuum, 3×10^{10} cm. per sec. and $t=0.137$ sec. Then

$$\frac{2\pi r}{\theta} \tan \theta = ct.$$

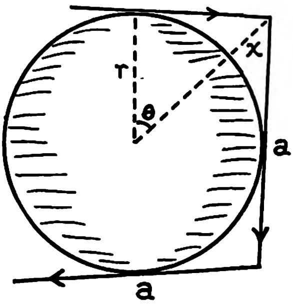


Fig. 2

Solving this for θ, gives $\theta=15$ deg. 1 min. Hence the polygon has 12 sides, and the height x of the corners above the surface of the earth is 142 miles.

This calculation is independent of group velocity considerations in the case of the wave polarized with electric vector parallel

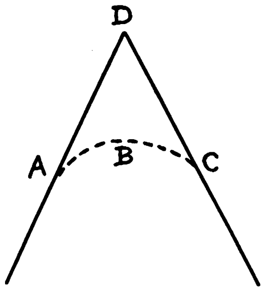


Fig. 3

to the magnetic field, for Breit and Tuve have shown[2] that for this wave the measured retardation along a curved path ABC (Fig. 3) is the same as that which would take place in vacuum along ADC where AD and DC are tangents at the points A and C.

[2] *Physical Review*, **28**, 554, 1926.

Thus the ray of Fig. 2 may be as in Fig. 4 with no change in the calculation. For other possible states of polarization the group velocity consideration may enter and may produce a small lengthening or shortening or displacement of the round-the-world pulse. Just to what extent this effect may exist will depend upon the distribution of the electrons in the upper atmosphere.

I do not know the accuracy of the time $t=0.137$ sec., but if it changes by $\pm$ 1 per cent, θ changes by $\pm$ 3 deg. (and x by $\pm$50 miles). It is seen that if θ does not have the value suited to the distances

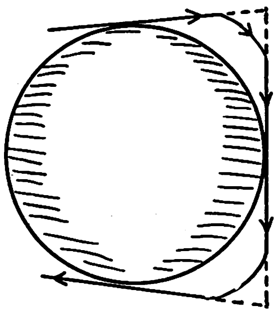


Fig. 4

between the transmitting station, the signal may be very weak. On this view small movements of the Kennelly-Heaviside layer would be expected to cause violent fading of the round-the-world signals (as is observed), whereas if the ray were as in Fig. 1 violent fading would perhaps not be expected. It also follows that a weak signal might have a time t different from 0.137 sec. Perhaps data may be obtained from which this might be decided.

Whether the round-the-world ray-path is that of Fig. 1 or Fig. 4 (or some other kind) depends upon the distribution of the electrons in the upper atmosphere. It is because of this that the question of the ray path is an interesting one.

MEASUREMENTS OF THE EFFECTIVE HEIGHTS OF THE CONDUCTING LAYER AND THE DISTURBANCES OF AUGUST 19, 1927*

By

Odd Dahl[1] and L. A. Gebhardt[2]

([1] Department of Terrestrial Magnetism, Carnegie Institution of Washington; [2] Naval Research Laboratory, Bellevue, Anacostia, D. C.)

Summary—An account is given of further improvements in the echo method of observing effective heights of the reflecting layer. A table of values is given showing effective heights at various times of day from August 15 to 25, 1927, covering a period of general disturbance in transmission phenomena. The table shows an increase of height after the disturbance as compared with the days preceding it. The data obtained are compared with those furnished by Mount Wilson on disturbances in sun spots as well as the general condition of radio reception. An unusually active spot was observed at Mount Wilson several days before August 19. If it was responsible for the disturbance its effect must have been cumulative. More systematic data are necessary to ascertain whether the rise in heights observed is characteristic for radio disturbances covering large areas.

General Method

The experiments described below were carried out during the summer of 1927. The method is that described by Breit and Tuve.[1] As before, the signals were transmitted at the Naval Research Laboratory and received at the Department of Terrestrial Magnetism.

The experimental arrangements were almost identical with those employed by Breit and Tuve. The general scheme was as follows: A crystal-controlled, 4,015 kilocycle, 10-kilowatt set located at Bellevue was modulated so as to send out short interrupted trains of waves. The duration of each train was roughly 1/1,500 second, the space between the end of one train and the beginning of another being about 1/750 second. The signals were received on the roof of the main building of the Department of Terrestrial Magnetism. The receiving aerial was a small loop. This was connected to the first detector of a superheterodyne set having an intermediate frequency of 50 kilocycles. The second detector of the superheterodyne was fed into a resistance of 25,000 ohms connected between the filaments and grids of four 7.5-watt

* Original Manuscript Received by the Institute, December 31, 1927.
* Presented before the convention of the International Union of Scientific Radiotelegraphy, October 14, 1927.

[1] G. Breit and M. A. Tuve, *Phys. Rev.*, vol. 28, 1926, pp. 554–575.

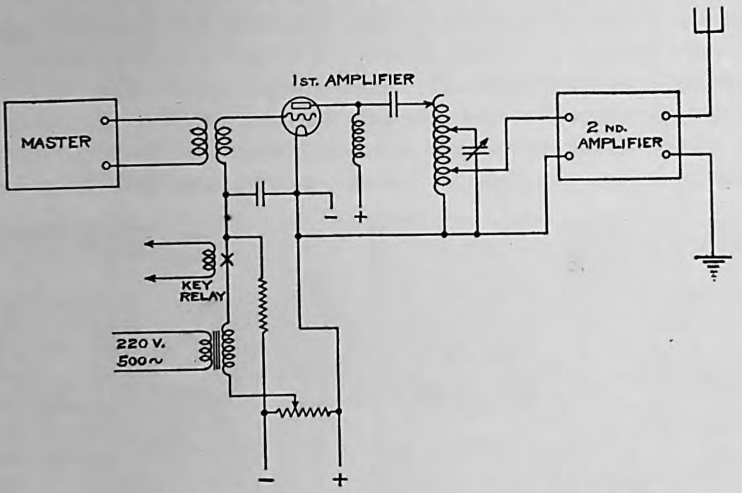


Fig. 1—Diagram of Transmitting Circuit, Crystal Control, 10 KW, 4,015 kilocycles.

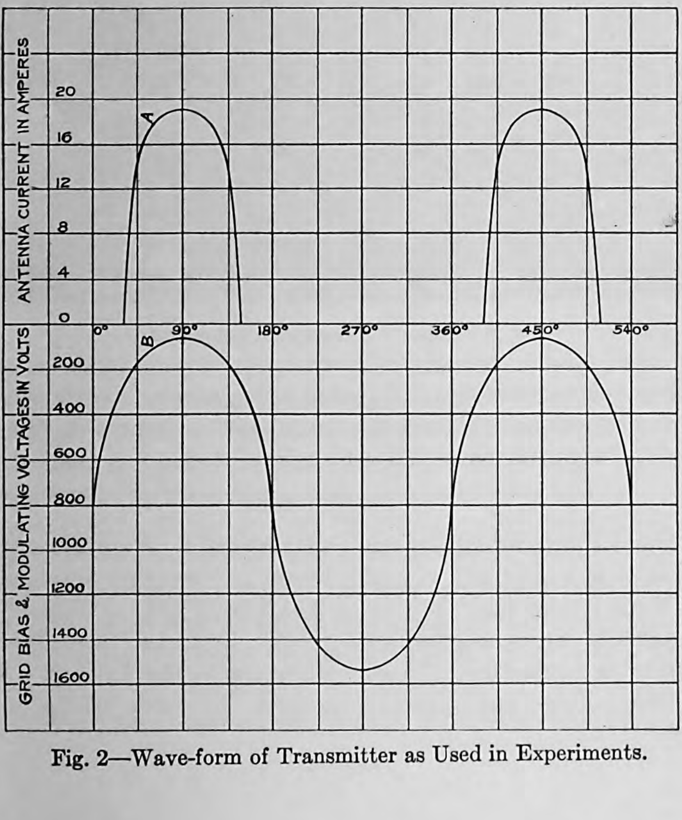


Fig. 2—Wave-form of Transmitter as Used in Experiments.

tubes (No. 210X) used in parallel. The output of these tubes was passed through a General Electric oscillograph with a proper balancing arrangement for the steady plate-current. The resultant wave-form was photographed and observed visually.

The idea of the method is to photograph on the same record the trains arriving directly over the ground from Bellevue as well

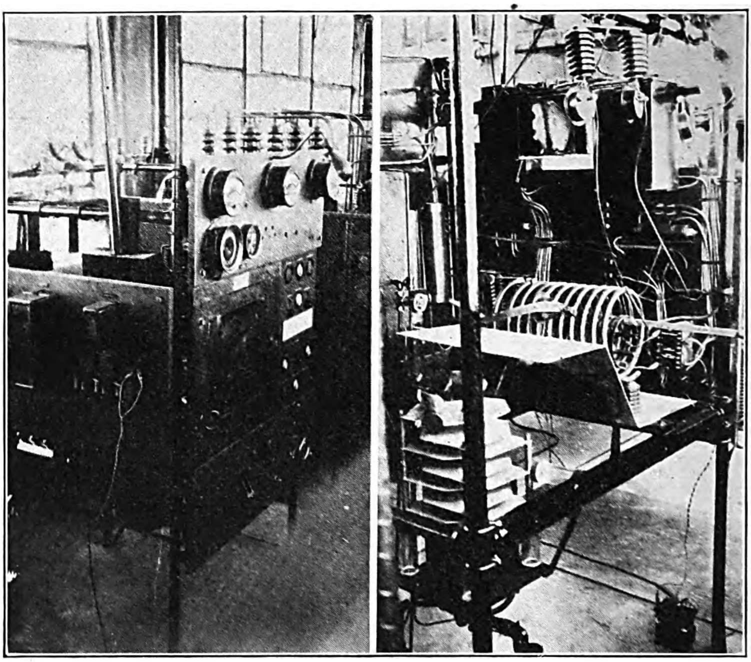


Fig. 3—Views of Transmitting Apparatus.

as their echoes arriving *via* the conducting layer. The lag between the arrival of the echo and the train which produced it gives the effective height of the layer.

Transmitter

Fig. 1 shows a diagram of the transmitter. A quartz crystal-oscillator was used as a master. Its output was first amplified by a 250-watt tube and then by a 20-kilowatt tube. Control and modulation were accomplished in the first, i.e., the 250-watt stage of the amplifier. In controlling, with the key up, a high negative voltage was placed on the grid. This stopped the ampli-

fying action of the 250-watt tube. The same principle was used in modulating.

The modulation voltage was obtained from a 500-cycle generator fed through a transformer. The 500-cycle output of the transformer was passed through a voltage divider, and a proper part of the grid biasing voltage was superposed on a constant negative bias. Fig. 2 shows the actual relations. The constant negative voltage was 800 volts. The alternating 500-cycle voltage has an

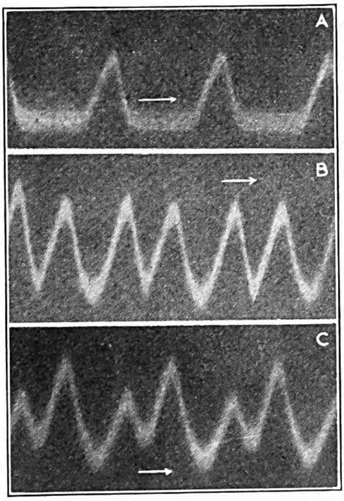


Fig. 4—Wave-forms Observed on August 25, 1927.
(A) Reflection absent, showing wave-form of transmitter. (B) Reflection approximately as strong as the ground-wave. (C) Reflection weaker than ground-wave.

effective value of 525 volts. Its graph against time is shown on curve *b*. The portions of time in which the 250-watt tube is operative are shown on *a*. The humps shown on *a* represent approximately the envelope of the antenna current.

Fig. 3 shows photographs of the transmitting apparatus.

Receiving Apparatus

The difference between the receiving apparatus used in these tests and in those of Breit and Tuve consists in the following items:

(1) Different location of the apparatus. The measurements were made on the roof of the main building, while in the summer of 1925 they were made in the small Experiment Building.

(2) The receiving aerial was a small loop 2 feet square instead of an antenna.

(3) 7.5-watt tubes were used in the power amplifier instead of the 5-watt tubes used previously. Four tubes in parallel were used in each case.

(4) A 7.5-watt oscillator loosely coupled to the superheterodyne was substituted for the 201A oscillator used previously.

Results

The results obtained are summarized in the following table.

Determination of Heaviside-Layer Heights

Date	Time	Height in miles	Transmitter
1927	h m h m		
Aug. 15	2:00– 2:15 P.M.	128 or 66	Antenna
16	10:00–10:15 A.M.	70 or 124	Antenna
16	2:00– 2:15 P.M.	62	Antenna
17	10:00–10:15 A.M.	60 and perhaps 110	Antenna
17	2:00– 2:15 P.M.	53	Antenna
19	....	No reflection	Antenna
22	2:00– 2:15 P.M.	112 or 80	Antenna
23	10:00–10:15 A.M.	103	Horizontal doublet
23	10:30–10:45 A.M.	104	Antenna
23	2:00– 2:15 P.M.	104	Antenna
23	2:30– 2:45 P.M.	124	Horizontal doublet
24	10:00–10:15 A.M.	92	Horizontal doublet (ground wave stronger than reflected)
25	10:15–10:30 A.M.	103	Horizontal doublet (reflected wave stronger than ground)

It must be noted also that not only did the reflected wave disappear on August 19 but also that the "ground" wave itself was received with difficulty. During the same period the reception of signals at the Naval Research Laboratory was decidedly below the average—some familiar stations not coming in at all.

Reports were received that the abnormal conditions caused a great deal of difficulty in copying on the San Francisco-Washington high-frequency circuit both day and night. For three days previous to and including the 21st much trouble was had in clearing traffic to London, it being the first time in over a month that London did not receive everything on the first transmission.

A report from London stated that during this period poor receiving conditions existed on all short-wave stations to the west of London, but that conditions were normal in receiving stations from other directions. Another London report stated

that the same trouble was experienced with the Canadian Beam. The conditions, however, apparently did not affect the reception of Washington's 12,045 kilocycles frequency in San Francisco. On August 19, Lakehurst, New Jersey, was unable to receive Washington's 8:15 a.m. weather broadcast on 4,015 kilocycles, and

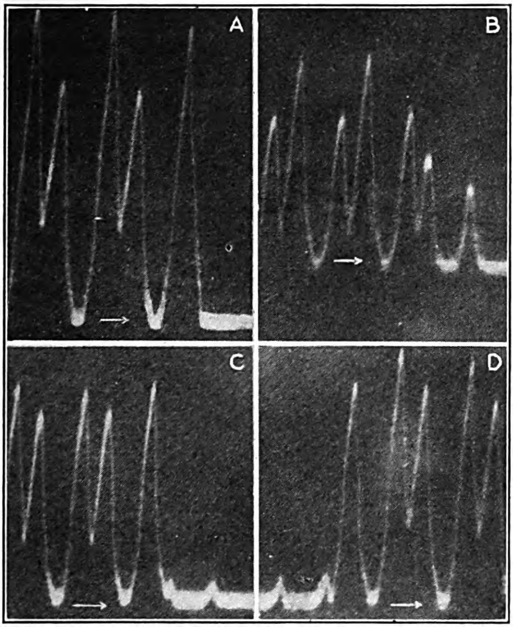


Fig. 5—Wave-forms Observed on October 8, 1927.
(A) End of dot, reflection bigger than ground-wave, no key-click. (B) End of dot, reflection smaller than ground-wave, last ground-hump weaker than preceding ones as also its reflection and in the right proportion. (C) End of dot, reflection slightly bigger than ground wave; note small key-click giving diminutive ground and reflected humps; time-separation between clicks same as between big humps. (D) Beginning of dot; key-click arrives first along ground, then as a reflection, then big ground-hump, its reflection, etc.; separation checks as in (C). Note that amplitude of ground-hump is constant for all.

Pensacola, Florida, could not copy this broadcast on 8,030 kilocyles. The failure of these two stations to get the broadcast was considered unusual, and could not be accounted for since the weather broadcast had been properly transmitted.

Several people reported to the Naval Research Laboratory that during the disturbance broadcast signals in the 550-1,500 kilocycles band also showed abnormalities.

We attempted to ascertain whether this unusual condition was definitely connected with a visible disturbance on the sun. No sufficiently definite information could be obtained because, even though there was an unusual spot in the southwestern quadrant of the sun, it had been active several days before August 19.

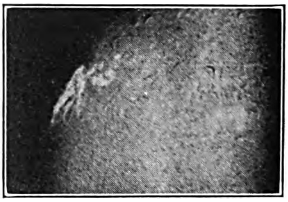


Fig. 6—H-alpha image of sunspot-group photographed at Mount Wilson, August 19, 1927, 14^h44^m Greenwich mean time.

However, our information seems fairly definite to the extent of indicating that just before the disturbance the layer height was around 50 miles, while after the normal condition had been reached again the layer height was increased to 100 miles and more.

Figs. 4 and 5 give reproductions of typical wave-forms observed during this period. The first of these was taken on August 25 and the second on October 8. Fig. 6 is a reproduction of the photograph of the sun on August 19 kindly put at our disposal by S. B. Nicholson of the Mount Wilson Observatory.

CORRELATION OF LONG WAVE TRANSATLANTIC RADIO TRANSMISSION WITH OTHER FACTORS AFFECTED BY SOLAR ACTIVITY*

By

Clifford N. Anderson

(American Telephone and Telegraph Company, New York City)

Summary—*Phenomena usually thought of as being affected by solar activity are: Sun spots, solar constant, earth's magnetic field, atmospheric electricity, auroras, earth currents, and, as has been recently shown, radio transmission. In order to correlate these with each other, the daily character of each must be reduced to some single figure. The factors which lend themselves most easily to quantitative correlation are sun spots, solar constant, earth's magnetic field, daylight radio transmission and night radio transmission. For sun spots, the numbers as prepared by Wolfer are used and for variations in the solar constant, the measurements by Dr. Abbot. Several sets of character figures of the earth's magnetic field were available. Of these, the three figures show too great a contrast between disturbed and undisturbed conditions while the Van Dijk show too little. A set of figures which showed a contrast between the disturbed and undisturbed conditions of the same order of magnitude as that obtained between disturbed and undisturbed radio conditions was devised. This method consisted of obtaining from the hourly averages, the total variation of the horizontal and vertical components of the earth's field. Such a figure is easily computed although better results might be obtained if the variation were taken directly from the magnetogram.*

The diurnal characteristic of long wave transatlantic radio transmission can be divided into four parts: daylight over entire transmission path, darkness over entire path, and the sunset and sunrise transition periods. Because the transition periods are normally characterized by one or more dips of signal field, some of short duration, they do not lend themselves particularly well to correlation. For daylight transmission, the average daylight fields were used. For night transmission, a character figure was derived for each day indicating the extent to which the night fields were reduced from their assumed undisturbed values. These assumed undisturbed values were obtained from the maximum night values and approached values given by the inverse distance law as their limit.

By mathematical analysis as well as by inspection high correlation was found between the general trends of these factors (except for the solar constant) which shows a decided decrease in 1926 while the other factors maintain their maxima. Weekly, three-weekly or quarterly averages of deviations from this trend show only a small correlation.

High daylight radio field strengths (at 57,000 cycles) obtain during periods of marked magnetic activity. In most cases, the magnetic disturbance precedes the high radio values, but there is evidence of the abrupt rise to high values preceding the magnetic disturbance and at times a gradual rise to high values independent of the magnetic activity. One interesting fact is the unusually low field strengths and

*Original Manuscript Received by the Institute, November 10, 1927.

low magnetic activity existing for several weeks on either side of December 1, agreeing quite well with the date of December 7 when the sun's equator (region of low solar activity) is on the line of centers of the sun and earth.

High disturbance of the night radio field (57,000 cycles) consistently occurs simultaneously with the magnetic disturbances but with a recovery lasting sometimes as long as six to eight weeks instead of the few days as is the case with the earth's field. If a second disturbance occurs before recovery is complete the effects are added to the existing state. Although fewer data are available on 17 kilocycles and 25 kilocycles, indications are that the maximum disturbance is not so immediate and the recovery sooner than on 57 kilocycles.

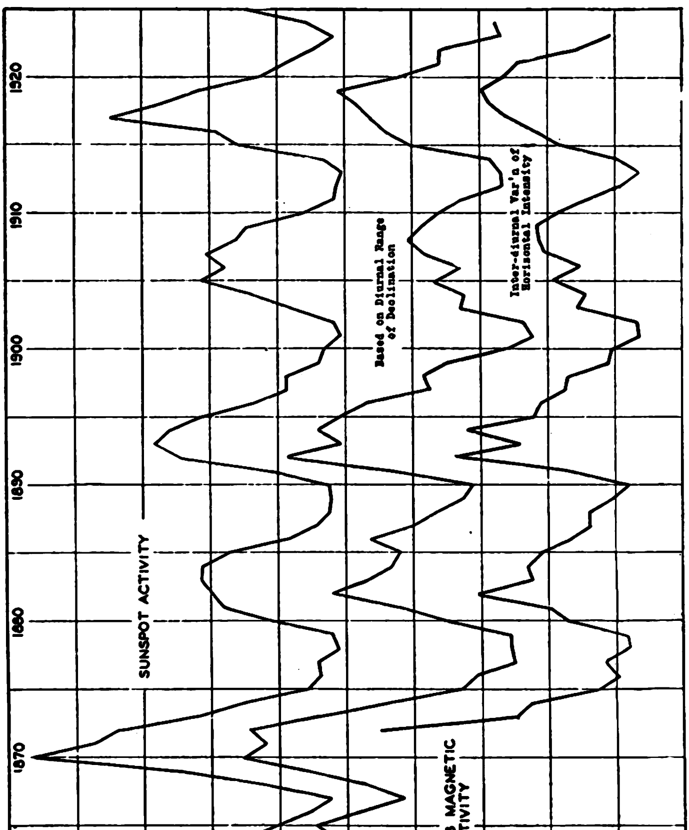
Qualitative correlation of disturbances on grounded telegraph lines indicates discrepancies in the times of beginning and in times of maximum disturbances in earth currents and earth's magnetic field which support conclusions reached by other investigators that these phenomena are not related in any simple way.

Such a statement in reference to all the phenomena examined is also appropriate as a general conclusion for this entire study.

I. Introduction

THE recent interest in the question of whether any significant relations can be established between variations in radio transmission and the activities of the sun has made it seem desirable to present in some detail the results of a study of long wave transatlantic radio transmission in its relation to a number of phenomena commonly thought to be manifestations of solar activity. In a paper presented before the Institute of Radio Engineers in May, 1925 (9), it was pointed out that there had been found a correspondence between times of abnormal radio transmission and the occurrence of severe magnetic storms. This is believed to be the first concrete evidence of an important relationship of this kind. Since that time not only has the transmission information obtained by the transatlantic radio telephone experiments been subjected to more careful examination, but also a considerable amount of additional information has been obtained.

In order to be reasonably certain that no possible factor of significance was being overlooked, it has seemed worth while to investigate these radio data in connection with data on the occurrence of sun spots, the variations of the sun's radiation (solar constant), disturbances of the earth's magnetic field, atmospheric electricity, the aurora, and earth currents. To carry out an examination of such interrelations as may exist on any simple basis it has seemed necessary to select for each of the phenomena some quantitative value which could be taken as representative of the conditions existing over an appreciable period of time such,



(Reproduced from *Terrestrial Magnetism and Atmospheric Electricity*, March, 1926.)

Fig. 1—The Activity of the Sun (sunspot numbers) and the Activity of the Earth's Magnetism as based upon absolute ranges of the magnetic declination at Kew and Cheltenham, 1856–1924, and upon the interdiurnal variability of the horizontal intensity at Bombay and Potsdam, 1872–1923.

for instance, as an entire day. To obtain such representative values as would be adaptable to the kind of analysis which has been made, it has been necessary to be somewhat arbitrary in choosing these "character figures." While an understanding and appreciation of the results of the entire study does not necessarily require

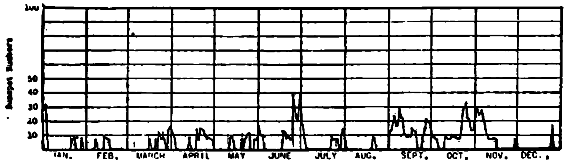


Fig. 2—Sunspot Numbers, 1923. These sunspot numbers are given by A. Wolfer, of Zurich, Switzerland, and published in *Journal of Terrestrial Magnetism and Atmospheric Electricity.*

an examination of the detailed information on which they are based, it has seemed desirable, not only for completeness, but as a possible aid to others interested in this kind of study and to those radio engineers who may wish to obtain an elementary acquaintance with the existing information relating to supposed solar phenomena, to include as a separate section of this paper the description of the available data and the methods which were

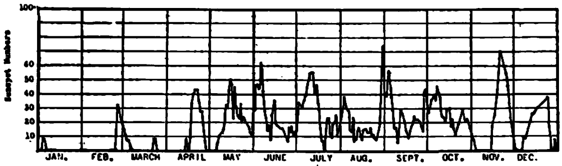


Fig. 3—Sunspot Numbers, 1924. These sunspot numbers are given by A. Wolfer, of Zurich, Switzerland, and published in *Journal of Terrestrial Magnetism and Atmospheric Electricity.*

used in bringing them into a form more easily handled. This material is included in Section II of the paper entitled "Description of Phenomena and Treatment of Data."

In the third section of the paper, entitled "Intercorrelation", is given the procedure of the study and a discussion of the results. The procedure was carried out along two general lines. The first was an examination of the information relating to the various phenomena intended to disclose whatever significant correspondences between abnormal conditions there might be. The second procedure was a strictly mathematical application of an intercor-

relation formula to determine whether there were any relations capable of being recognized by the limited analysis which this formula permits. While this intercorrelation formula is capable of displaying in a quantitative and summarized manner the ex-

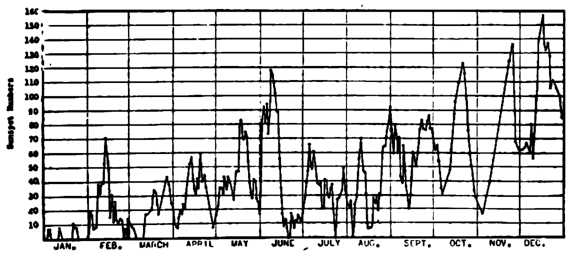


Fig. 4—Sunspot Numbers, 1925. These sunspot numbers are given by A. Wolfer, of Zurich, Switzerland, and published in *Journal of Terrestrial Magnetism and Atmospheric Electricity.*

tent to which any two sets of data are related in their point-by-point variations, it does not discriminate between differences in the duration of the disturbances in the two phenomena. Since the inspection of data disclosed many cases where apparent significant correspondences of abnormal conditions had a short period

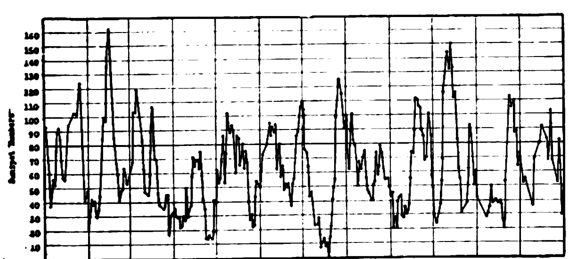


Fig. 5—Sunspot Numbers, 1926. These sunspot numbers are given by A. Wolfer, of Zurich, Switzerland, and published in *Journal of Terrestrial Magnetism and Atmospheric Electricity.*

of existence for certain of the factors but a relatively long period of recovery for others of the factors, it was not to be expected that the mathematical intercorrelation would give results of great significance for relatively short intervals of time. It is, however, capable of sifting out from a set of information which to the eye appears entirely random, general relationships which are of value.

The results of the study may be briefly abstracted as follows: The great reduction in night-time signal fields occurs essentially

at the same time that a disturbance in the earth's magnetic field occurs. However, instead of reverting to normality within a day or so as is the case with the earth's field, the recovery of the high night radio fields is very gradual, sometimes requiring six to eight weeks before completion. If no allowance be made for this difference in the recovery, the mathematical correlation of day-to-day values of the two phenomena is not particularly large.

High daylight radio fields accompany periods of high solar activity rather than individual storms. This is, in general, true

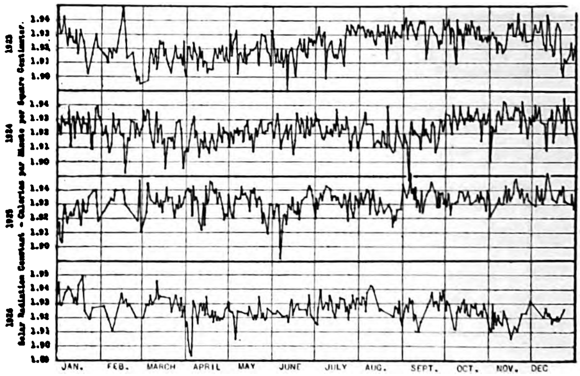


Fig. 6—Daily Variation of Solar Radiation Constant. Data taken by C. G. Abbot and colleagues of the Smithsonian Institution. These measurements represent means of measurements made at Harqua Hala, Arizona, and Montezuma, Chili. Measurements are made with a silver disk pyrheliometer.

of the intercorrelation of other phenomena associated with solar activity. In fact mathematical correlations of radio with those phenomena such as sun spots, disturbances in the earth's field and variations in the solar constant are higher than the corresponding correlations of these phenomena between themselves. Such correlations lie mainly in the broad movement corresponding to the 11-year cycle of solar activity and that with day-to-day, week-to-week, or even three-month averages of deviations from this trend the correlations are small.

Disturbances on grounded telegraph lines also occur chiefly during periods of high solar activity and although they usually accompany severe magnetic storms, discrepancies in the times of beginning and the times of maxima indicate that these phenomena are probably not related in any simple way.

The radio transmission phenomena associated with sunrise and sunset are of interest.

II. Description of Phenomena and Treatment of Data

Sun Spots

As it has been found in general that sun spots, solar constant, variations in earth's magnetic field, potential gradients of the atmosphere, earth currents, auroras or radio vary more or less together in periodic cycles, they have been collectively assumed

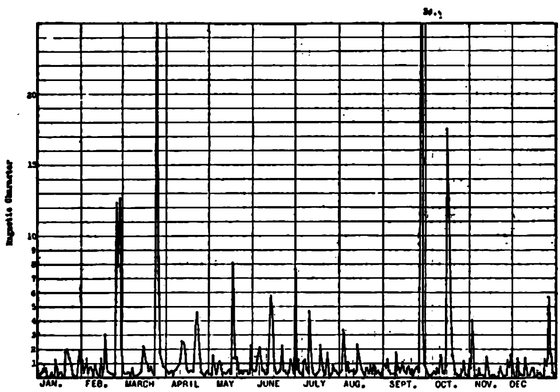


Fig. 7—Magnetic Character, Chree, 1923. These magnetic characters above are the sum of the hourly magnetic characters (X 10^{-4}) for each day as computed. The characters are proportional to the sum of the squares of extreme ranges of variation of declination and horizontal and vertical components of the earth's field. Because of the second power, they over-emphasize the disturbed days.

to denote the states of "solar activity." Of all the indices of solar activity, the observations of sun spots are probably most direct in that the measurements are made directly on the sun, weather conditions permitting.

Sun spots seem to be similar in nature to cyclonic areas which occur in the earth's atmosphere. The spots appear to consist of funnels of ascending gas, which cool on expanding at the higher levels and so appear darker than the surrounding vapors. They develop quite suddenly. Although they do have some motion of translation of their own, their movement as viewed from some fixed point in space is, in general, that due to the rotation of the sun about its axis. The period of this rotation varies with latitude on the sun, being 24.5 days at the sun's equator and 31 days

at a latitude of 80 deg. Sun spots rarely occur at the equator or more than 35 deg. from the equator; and as the majority occur in the band from 10 deg. to 30 deg. on either side of the equator, the period is approximately 25 to 26 days. Allowing for the movement of the earth on its orbit, the average period is about 27.2 days. Their length of life is variable, ranging from a matter of hours to that of months. One has been recorded as having stayed in the same place on the sun's surface for 18 months.

Over a larger period of time, the number of sun spots varies in periodic cycles averaging 11.2 years. Since 1788 the interval

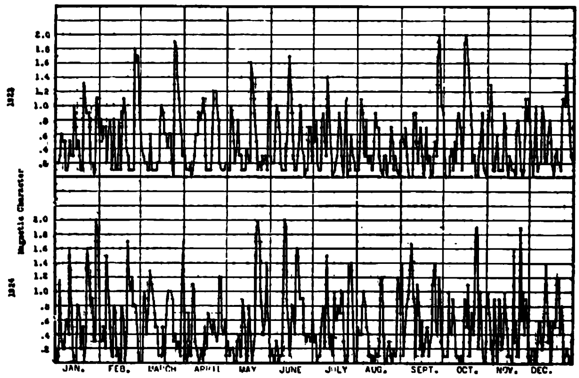


Fig. 8—Magnetic Character (Van Dijk), 1923-1924. The magnetic figures above are those given by Dr. Van Dijk and published in the *Journal of Terrestrial Magnetism and Atmospheric Electricity*. They represent results obtained at several observatories throughout the world. The figures are estimated from the character of the magnetograms. Because the maximum figure is 2 the tendency is not to give severe disturbances proper weight.

between maxima has varied, however, from 7.3 to 17.1 years. This periodic activity from 1856 to 1924 is shown in Fig. 1 (1). It has been suggested that the periodic occurrence of sun spots may be a tidal effect produced by the planets, of which Jupiter's influence predominates. Jupiter's period is 11.86 years.

The sunspot numbers referred to in this paper are those originally prepared by Wolf and continued by Wolfer. They are based in part on the number of spots and in part on their size. The formula used is $r=k\ (10\ g+f)$ where g is the number of groups and single spots observed, f the total number of spots which can be counted in these groups and single spots combined, and k a multiplier which depends on the conditions of observation

and the telescope employed. The sunspot numbers cannot take into account sun spots on the sun's hemisphere away from the earth or disturbances, indications of which might not have reached the surface. The sunspot numbers for 1923 to 1926, inclusive, are shown in Figs. 2, 3, 4, and 5. These data extend essentially from a period of minimum sunspot activity to a period of maximum activity.

In the same category with sun spots are measurements of flocculas, faculas, solar prominences, magnetic fields of sun spots, etc.

Solar Constant

Another method which makes measurements of the sun itself is that of measuring the Solar Constant. This so-called *constant*

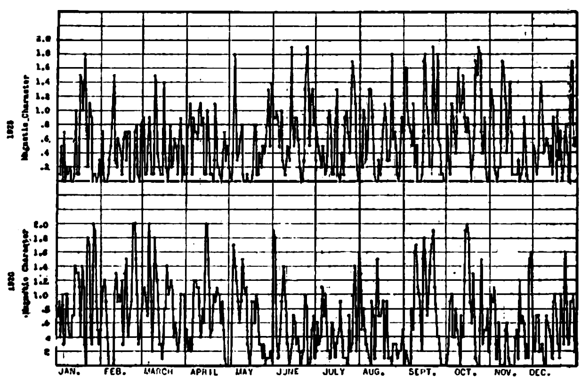


Fig. 9—Magnetic Character (Van Dijk), 1925-1926. The magnetic character figures above are those given by Dr. Van Dijk and published in the *Journal of Terrestrial Magnetism and Electricity.* They represent results obtained at several observatories throughout the world. The figures are estimated from the character of the magnetograms. Because the maximum figure is 2 the tendency is not to give severe disturbances proper weight.

is the energy received from the sun in calories per minute per square centimeter at the earth's surface assuming no atmosphere and that the earth is at its mean distance from the sun. Recent figures indicate the average to be about 1.94 calories. The extreme range of variation about this mean over a period of years is only a matter of $\pm$ 2½ per cent. High solar radiation accompanies high solar activity. In addition there are superimposed periodic fluctuations of 25⅔ months, 15 months and 11 months. Of these the first is the strongest. Most of the variation is localized in the

ultra-violet region of the spectrum with a day-to-day range at 0.29 micron of probably as much as 100 per cent.

The measurements are made by Dr. C. G. Abbot and his colleagues of the Smithsonian Institution at Montezuma, Chili, and Harqua Hala, Arizona. The method is that of the silver disk pyrheliometer in which a silver disk is exposed to sunlight for several periods of 100 seconds each and the temperature rise noted. Knowing the thermal constants of the apparatus, the received energy can be computed.

The outstanding difficulty of this method is no doubt that of eliminating the effect of the atmosphere. Many of the radiation

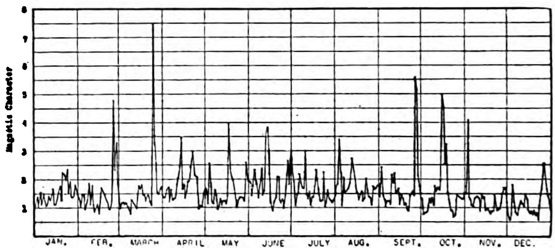


Fig. 10—Magnetic Character (Total Daily Variation). Data taken at Cheltenham, Md., U. S. Coast and Geodetic Survey, 1923. The magnetic characters as represented above are determined by adding the total variation between hourly averages of horizontal and vertical field (gammas) for each day and dividing by 100.

components never reach the earth's surface; and all are absorbed to a certain extent. Corrections are made for haziness and water vapor. The method does not admit of correction for volcanic dust, etc., and in general is satisfactory only for stations of excellent and uniform conditions. The data for the years 1923 to 1926 are shown in Fig. 6 (4).

Earth's Magnetic Field

General. As is well known, the earth acts like a magnetized sphere whose negative pole is near Boothia Felix approximately 71 deg. N. latitude and 96 deg. W. longitude and whose positive pole, although it has never been reached, is about 73 deg. S. latitude and 156 deg. E. longitude. At the earth's surface at Cheltenham, Maryland, the horizontal component is approximately 18,800 gammas (1 gamma = 0.00001 gauss) and the vertical component approximately 55,000 gammas. In equatorial regions the field is

mainly horizontal and in the polar regions it is chiefly vertical. Because of the fact that the magnetic and geographical poles do not coincide, the horizontal magnetic field is not parallel to the geographical meridians but differs by an angle called the declination which varies for various geographical locations. For this reason, the horizontal field is sometimes divided into two components, the north (X) and the west (Y).

There is a daily variation in the various magnetic components which is a local phenomenon depending upon the altitude of the sun, the field being usually quite constant during the night. In the United States in summer, the compass needle points 10 min.

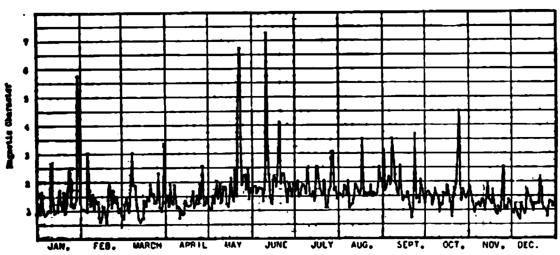


Fig. 11—Magnetic Character (Total Daily Variation). Data taken at Cheltenham, Md., U. S. Coast and Geodetic Survey, 1924. The magnetic characters as represented above are determined by adding the total variation between hourly averages of horizontal and vertical field (gammas) for each day and dividing by 100.

more to the west at 1 P.M. than at 8 A.M. The variation of horizontal and vertical components changes with latitude. At stations within 25 deg. or 30 deg. of the equator, the field *increases* during the morning to a maximum shortly before noon, after which it again decreases. For latitudes above 40 deg., the field *decreases* to a minimum shortly before noon. The normal range therefore depends upon the place of observation at Cheltenham, Maryland, being in the order of 0.1 per cent to 0.2 per cent.

From year to year the field drifts, sometimes increasing and sometimes decreasing. For example, the decrease in the total field at Cheltenham, Maryland, from 1905 to 1924 has been of the order of 1800 gammas or about 0.3 per cent of the total field.

The earth's field is also subject to disturbances of varying degrees, the range, however, for even the "violent" disturbances being for the middle latitudes only of the order of 1 or 2 per cent, although cases have been recorded involving a change in the horizontal component of as much as 5 per cent. The amplitude of

the fluctuations is greater in high latitudes than in equatorial regions. There is no difference in intensity as regards the day and night hemisphere, although there seems to be some indication that the P.M. hemisphere is slightly more disturbed than the A.M. hemisphere.

In general, disturbances are characterized by rapid and excessive fluctuation in the earth's field with a net increase in the vertical field and a decrease in the horizontal field. These disturbances begin simultaneously, to within a minute over the whole earth's surface and are most intense the first day of the storm.

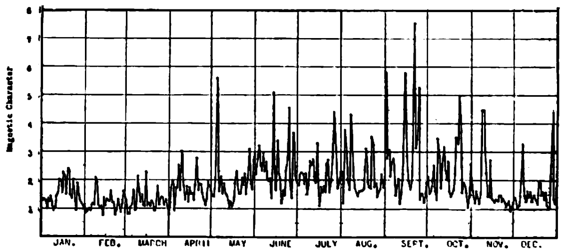


Fig. 12—Magnetic Character (Total Daily Variation). Data taken at Cheltenham, Md., U. S. Coast and Geodetic Survey, 1925. The magnetic characters as represented above are determined by adding the total variation between hourly averages of horizontal and vertical field (gammas) for each day and dividing by 100.

Then they gradually subside, reaching normality at the end of three or four days.

Theories. No proved theory has been developed to account for the existence and variation of the earth's magnetic field. Mathematical analyses indicate that about 96 per cent of the total field is due to internal forces, part to the vertical atmospheric conduction current and 2 to 3 per cent may be due to forces outside the earth. This latter amount is equivalent to approximately the total disturbance occurring at times of so-called magnetic storms. The correlation of magnetic storms and sun spots and the diurnal variation in the earth's field dependent upon local sun time rather than universal time suggest that the sun is responsible for part of the field. Practically all great magnetic storms occur simultaneously (to within a minute) over the entire earth; and because of the tendency to recur every 27 days, the explanation has been advanced that the storms are due to more or less sharply defined streams, at least on the advance front, of electrons emitted from

disturbed localities on the solar surface. The angular velocity, of the sun is such that the velocity of a stream relative to the earth would be approximately 400 km. per second and a radius of the sun would cross the earth's disk in about 30 seconds. Assuming a disturbance with a duration of one day, the width of the stream would be of the order of 35,000,000 km.

An explanation given for the greater frequency of magnetic disturbances during the equinoxes as compared with the solstices is the relative positions of the earth and the solar latitude of greatest sunspot frequency and greatest latitude of assumed solar activity. The solar equator is inclined 7 deg. 15 min. to the

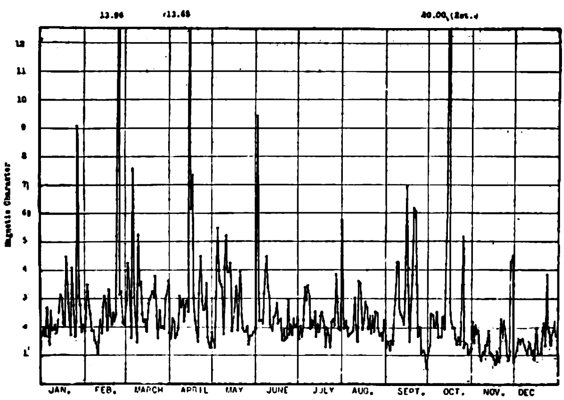


Fig. 13—Magnetic Character (Total Daily Variation). Data taken at Cheltenham, Md., U. S. Coast and Geodetic Survey, 1926. The magnetic characters as represented above are determined by adding the total variation between hourly averages of horizontal and vertical field (gammas) for each day and dividing by 100.

plane of the ecliptic as compared with 23½ deg. in the case of the earth. A line from the sun's center to the earth passes through the sun's equator on June 6, and December 7, and the maximum angle of 7 deg. 15 min. between that line and the equator occurs on September 8, and March 6.

The stream theory alone is not sufficient to account for the magnetic currents, but recourse has to be made to the idea of electric currents flowing in a conducting layer. One such theory (10) calls for a flow from west to east during the first phase of a storm when the horizontal field is slightly increasing and a current from east to west when the intensity of the storm increases and the horizontal field decreases. The interaction

between these currents and the horizontal field of the earth is presumed to result in a vertical movement downward during the first phase of the storm and later an upward movement.

Measures of Magnetic Activity. Various schemes have been proposed for giving a single figure indicating the magnetic character of a day. A rough indication is given by dividing days into three classes designated respectively by 0, 1, and 2; 0 denotes quiet conditions; 2, severely disturbed; and 1, intermediate conditions. This is a very rough way of estimating. The results depend largely upon the judgment of the observer and may indicate different things during periods of great or little activity.

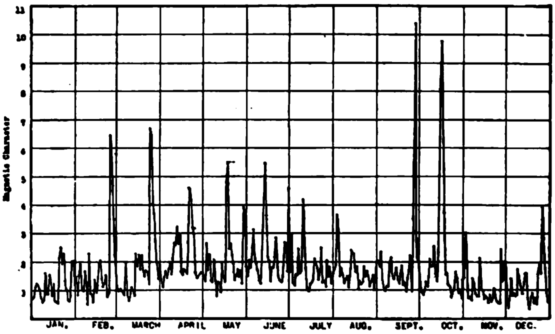


Fig. 14—Magnetic Character (Total Daily Variation). Data taken at Eskdalemuir, Scotland, 1923. The magnetic characters as represented above are determined by adding the total variation between hourly averages of horizontal and vertical field (gammas) for each day and dividing by 100.

Other methods for determining magnetic character figures have been devised and are given below—

Chree—Kew Observatory $R^2=\frac{1}{100}(R_D^2+R_H^2+R_Z^2)$

Schmidt—Potsdam Observatory $A=A_D+A_H+A_Z)$

Van Dijk—De Bilt Observatory $R=(R_D+R_H+R_Z)$

Bauer—Carnegie Institute HR_H

where R = Absolute diurnal range or difference between extreme daily values.

A = Range of the hourly mean values.

H = Horizontal intensity

Z = Vertical intensity

D = Declination

The character figures obtained by the Chree formula were available for each hour of 1923. The total of the 24 figures for each day (proportional to arithmetic mean with constant base line) is plotted in Fig. 7. They could perhaps more justifiably be combined in some other way but this way will serve to indicate quite well the nature of the figures derived by the three formulas. As the character figures are proportional to the sum of the squares of extreme ranges of variation of declination and hori-

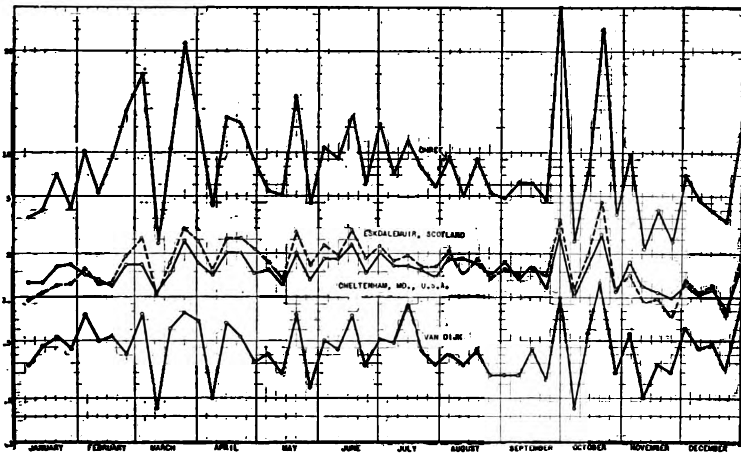


Fig. 15—Comparison of Weekly Averages of Magnetic Character Figures, 1923. This shows the good agreement between magnetic character figures by applying the total variation method to data from the observatories at Cheltenham, Maryland, and Eskdalemuir, Scotland.

zontal and vertical components, they over-emphasize the disturbed days.

The Van Dijk character figures as given in Figs. 8 and 9, are not computed by the formula attributed to Van Dijk in the table above but are estimates made by Dr. Van Dijk of the character of the magnetograms obtained from some 40 observatories throughout the world. Because the maximum figure is 2, the tendency is not to give severe disturbances proper weight.

The sum of the hourly variations for the 24 hours of the horizontal and vertical components has been found to be a convenient measure. As only the hourly means were available, the absolute total variation is, of course, greater in each case and such a measure determined from the magnetograms might be more significant. The computation is simple and for reasons shown later

seems to have some value. The data for the years 1923-1926 inclusive are shown in Figs. 10 to 13, inclusive. The unit might be designated:

$$V=\frac{1}{100}(V_H+V_Z)$$

This has been found to be somewhat better than the other magnetic character figure for comparison with radio transmission.

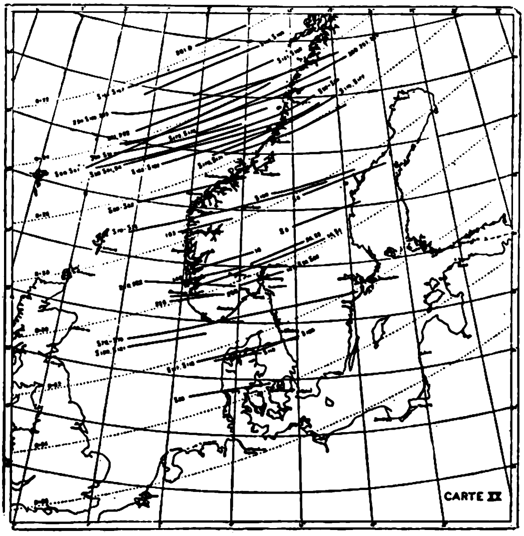


Fig. 16—Coincidence of Auroral Arcs with Magnetic Parallels.

Fig. 14 shows the above method applied to magnetic data obtained at the observatory at Eskdalemuir, Scotland. Comparison with Fig. 10 shows very good agreement indeed between disturbances in the earth's field on the two sides of the Atlantic. The character figures are computed for days beginning midnight to midnight local time so that if allowance were made for the 5-hour difference, the agreement would be even better. There seems to be a tendency for greater fluctuation in the field as measured in Scotland, and this is shown a little better in the weekly averages of Fig. 15. This figure shows also the Van Dijk and the Chree character figures for the same year. On account of the high correlation between the Scotland and the American data, no great error is introduced by using magnetic data at only one end of the

circuit. Computed correlation figures show that the "total variation method" agrees better with each of the Chree and Van Dijk sets of figures than the two latter do between themselves. This is interpreted as meaning that the results obtained by the total variation method lie between those obtained by the other

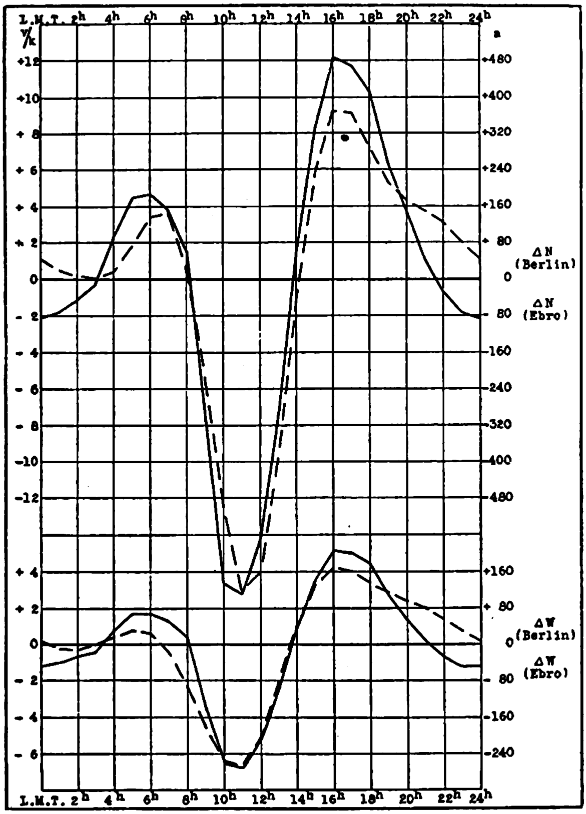


Fig. 17—Diurnal Variation of Earth-Current Components for the Ebro Observatory, 1914–1918, and for Berlin (Weinstein data, 1884–1887).

two methods. The Chree and Van Dijk figures have, however, been useful in checking conditions.

Atmospheric Electricity

Measurements of atmospheric electricity made at the bottom of the atmospheric sea are so affected by local conditions and factors unrelated to solar activity that it is difficult to draw con-

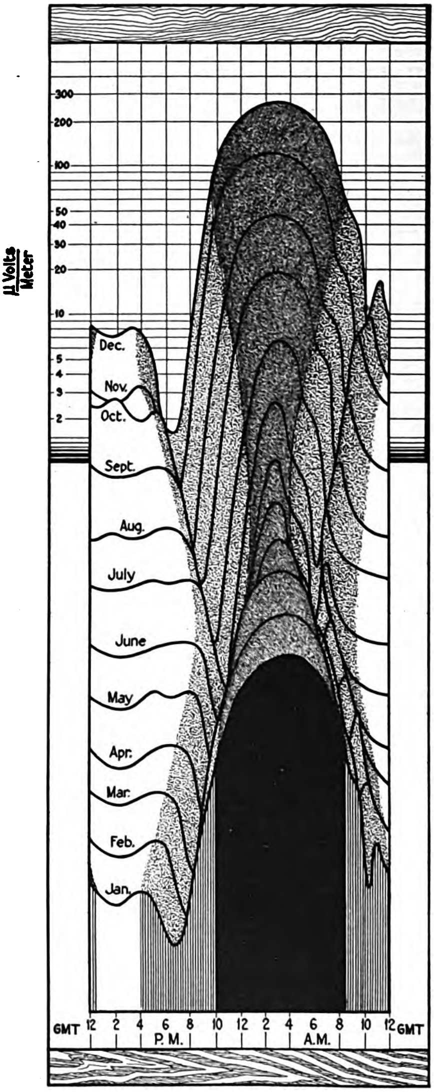


Fig. 18—Transatlantic Radio Telephone Transmission. Diurnal variation of assumed undisturbed signal field strength transmission. Rocky Point, Long Island to New Southgate, England. 5480 km., 57,000 cycles.

clusions as to the fundamental variations and as to what the effects of solar disturbances are.

It is well known, however, that the earth is negatively charged with respect to the surrounding atmosphere, the potential gradient at the earth's surface being of the order of 100 volts per meter. The value decreases with altitude, as indicated by balloon measurements; and it is believed that above 10 km. the potential becomes uniformly of the order of 1,000,000 volts positive with respect to the earth. This field tends to drive negatively-charged particles upward and positively-charged particles downward, and from the conductivity of the atmosphere and the potential gradient the current which flows from the atmosphere into the ground can be computed. This amounts to approximately 2 microamperes per square kilometer, or of the order of 1000 amperes for the entire earth's surface. This value holds for several kilometers above the earth because, although the gradient is decreasing, the conductivity is increasing. (At 9.5 km. Kolhörster found the ionization had increased tenfold.)

Why the earth's charge is not soon neutralized by this current is not known. It is thought that this current might be compensated for by:

(a) Convection currents of charged air
(b) Rain (negatively charged)
(c) Upward conduction currents where potential gradient is negative. (During thunder storms the potential gradient often exceeds 10,000 volts per meter and may be either positive or negative).
(d) Radiations received from outside: β rays from sun to which auroras are attributed might cause a negative potential gradient in vicinity of poles (See "C"); or cosmic rays of

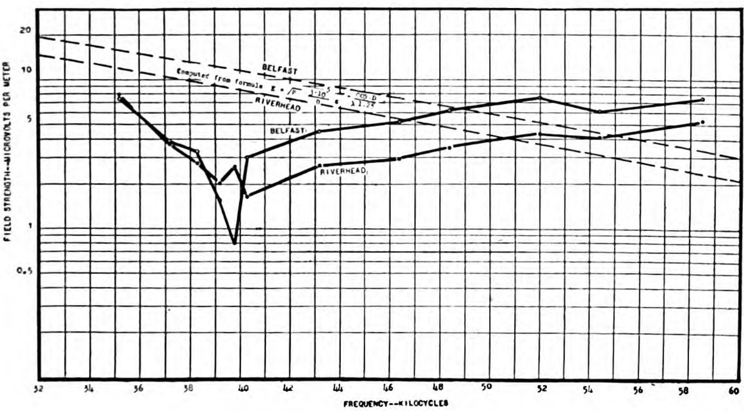


Fig. 19—Transatlantic Radio Telephone Transmission. Variation of daylight signal field with frequency transmission from Northolt, England (GKB)—2 kw. radiated power. Measurements made July to October, 1925.
Data on 58.5 kc. taken October 4, when, due to solar disturbances, field strengths were approximately twice those obtaining on the same frequency in July.

high penetration 15,000,000 volts as found by Hess, Kolhörster, Millikan and others, dissipating their energy in the production of β rays at the earth's surface.

In general the potential gradient decreases with the altitude of the sun, being lower at midday, in the summer time, and in the low latitudes (polar regions also low). In addition there seems to be in the diurnal variation, a variation not dependent on local time which consists of a minimum about 0400 GMT and a gradual increase to a maximum about 1800 GMT. The difference between land and sea values is small.

The correlation between variations of atmospheric electricity as measured at the earth's surface and other solar evidences is not so evident as it might be. However, for stations quite free

from meteorological disturbances there appears to be in general an increase in the potential gradient with increasing solar activity.

AURORAS

Auroral displays are closely connected with other phenomena usually associated with solar activity. The displays take various

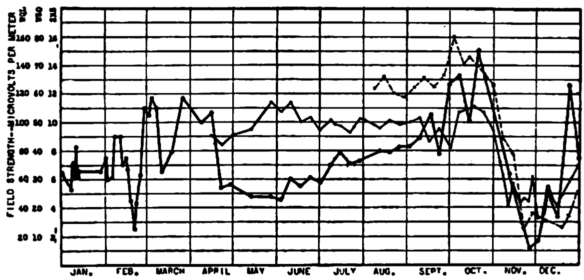


Fig. 20—Variation of Average Daylight Radio Field Strength, 1923.
———Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km., 300 amperes. Antenna current, 57 kc.
———Rocky Point, L.I. (WQL) to New Southgate, England, 5480 km., 600 amperes. Antenna current, 17.1 kc.
-------Marion, Mass. (WSO) to New Southgate, England, 5280 km., 600 amperes. Antenna current, 25.7 kc.

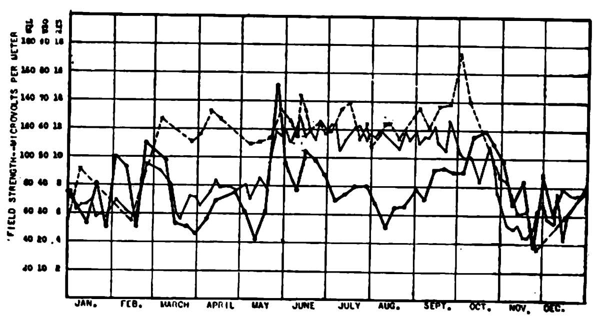


Fig. 21—Variation of Average Daylight Radio Field Strength, 1924.
———Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km., 300 amperes. Antenna current, 57 kc.
———Rocky Point, L.I. (WQL) to New Southgate, England, 5480 km., 600 amperes. Antenna current, 17.1 kc.
-------Marion, Mass. (WSO) to New Southgate, England, 5280 km., 600 amperes. Antenna current, 25.7 kc.

forms such as foggy auroras, pulsating auroras, diffuse arcs, split bands, drapery-shaped arcs, draperies, rays, corona and intermediate types. The ray form is probably the structural element of the other forms. They appear most frequently in the polar

regions with their distribution symmetrical with the magnetic axis rather than with the geographical axis. The isochasm (curve indicating same number of auroras) of maximum auroras is approximately 23 deg. from the intersection of the magnetic axis with the earth's surface. As this point is approximately 78 deg. N and 69 deg. W or not far from Etah, Greenland, the line of minimum aurora passes through the northern point of Norway (latitude 70 deg.) just south of Point Barrow, Alaska, (latitude 70 deg.) middle of Hudson Bay (latitude 60 deg.) and south of Greenland (latitude 58 deg.). In the United States auroras are seldom observed below the 30 deg. parallel and in Russia seldom

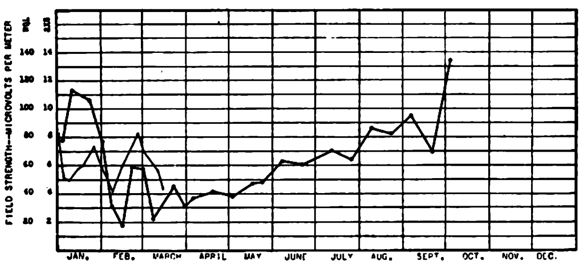


Fig. 22—Variation of Average Daylight Radio Field Strength, 1925. ———Rocky Point, L.I. (2XS) to Chedzoy, England, 5300 km., 300 amperes. Antenna current, 57 kc. ———Rocky Point, L.I., (WQL) to Chedzoy, England, 5300 km., 600 amperes. Antenna current, 17.1 kc.

below the 50 deg. parallel. The more severe the disturbance, the lower is the latitude at which the aurora is visible. Stormer reports that all the auroras observed at Oslo, Norway, during the years 1911–1922 were accompanied by magnetic disturbances with Van Dijk magnetic character figures of 2 with one exception—that of April 18, 1917, which had a character figure of 1. While the auroras observed at the lower latitudes follow in general the severe magnetic storms (most frequent during the equinoxes), the auroras observed near the auroral zone are most frequent in midwinter.

The more severe the disturbance, the greater are the higher and lower limits to which the aurora extends. By means of simultaneously photographing auroras against a background of stars from two observatories in the order of 30 km. apart, Stormer, Vegard, and Krogness (7) were able to determine the heights of auroras. In general the lower edge of the aurora occurred most frequently at an altitude of about 100 km. to 110 km., although at

times as low as 80 km. and as high as 200 km. or even, in exceptional cases, 400 km. The upper part of the aurora usually fades away to invisibility so that generally a well-defined upper limit does not exist. Indications are, however, that the average upper limit of diffuse arcs is of the order of 140 km., draperies 175 km., and rays 230 km. It is not unusual for the latter to extend much higher, even up to 800 km. (March 22–23, 1920).

It was also found that the direction of the auroral arcs and bands coincide approximately with the magnetic parallels, i. e., at right angles to the earth's field. The angle is not, however, exactly 90 deg. as south of the auroral zone the western end is 10

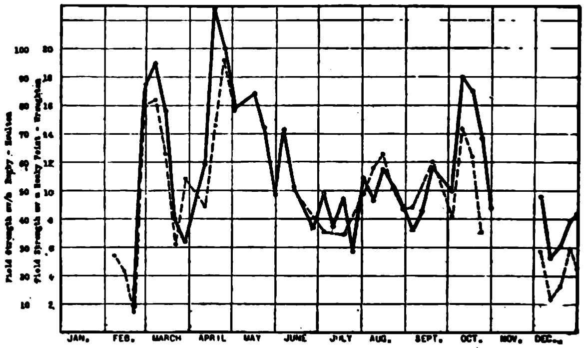


Fig. 23—Variation of Average Daylight Radio Field Strength, 1926. ———— Rocky Point, L.I. (2XS) to Wroughton, England, 5385 km., 300 amperes, Antenna current, 57 kc.
-------Rugby, England (GBT) to Houlton, Maine, 4670 km., 250 amperes. Antenna current, 57 kc.

deg. north of the normal to the magnetic meridian. (Fig. 16) (8). Inside the auroral zone the directions become less constant and at the magnetic axis point all directions are equally probable. There is also a small diurnal variation in the mean direction of the arcs, the angle decreasing from evening to morning.

Birkeland assumed the auroral illuminosity to be caused by the bombardment of the rarefied atmosphere by negatively charged corpuscles from the sun. By exposing a magnetized sphere to cathode rays he was able to produce a similar phenomenon. Experiment and mathematical theory both show that the electron streams should follow curved paths. These paths will not, therefore, coincide with magnetic parallels, but deviate from them by a small angle. Assuming the charge to be positive, the deviation is in the opposite direction to that observed.

Earth Current

General. Another factor which has been found to accompany solar disturbances is that of earth currents. When a conductor is grounded at two separate points a current will flow, indicating a difference in potential between these two points. There is not a great amount of data available on earth currents, but indications are that there is always a current flowing in a greater or less degree and that it is affected at times of unusual solar activity.

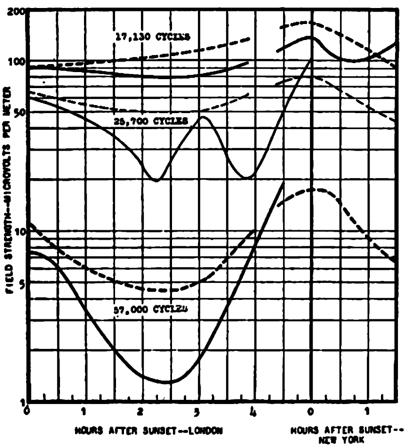


Fig. 24—Major Sunset Effects on Radio Transmission. Transmission from Rocky Point, L. I. (57 kc. to 17.1 kc.) and Marion, Mass. (25.7 kc.) to New Southgate, England. Solid line indicates undisturbed conditions, and dashed line disturbed conditions.

Measurements made in Spain by the Ebro Observatory at Tortosa indicate the current there to flow from a direction 29 deg. west of north with the north-south component of the gradient 0.2 volt per kilometer and the west-east component 0.11 volt per kilometer. The resultant, 0.23 volt per kilometer would increase to 0.8 to 1 volt per kilometer during electric or magnetic storms. As a relatively large part of the residual current is due to the electrochemical action at the contact between the electrode and the soil, the figures are to be taken only as indicating the approximate magnitude of the values involved. Fig. 17 shows the diurnal deviation from the mean of earth currents as measured at the Ebro, Spain, observatory and on telegraph lines between Berlin and Dresden (11). The minimum is seen to occur shortly

before noon, a primary maximum in late afternoon and a secondary maximum in early forenoon. The diurnal variation is more severe along a meridian (north-south) than at right angles to it (east-west). It is maximum during the equinoctial months and lowest near the solstitial months and increases with increase in sunspot activity. A careful comparison, however, with the earth's magnetic field, as measured at the same observatory, indicates time phase and directional discrepancies, from which it is concluded that they are related in only a minor way and probably only through a common cause.

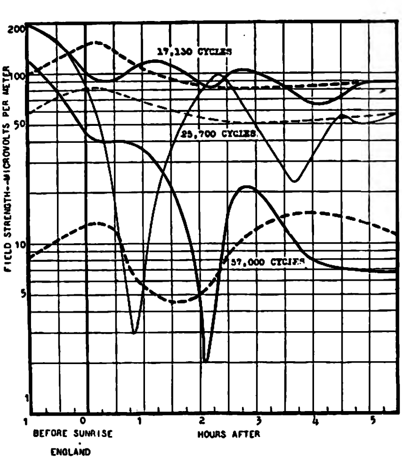


Fig. 25—Major Sunrise Effects on Radio Transmission. Transmission from Rocky Point, L. I. (57 kc. and 17.1 kc.) and Marion, Mass. (25.7 kc.) to New Southgate, England. Solid line indicates undisturbed conditions and dashed line disturbed conditions.

Earth Currents on Telegraph Lines. On days of severe solar disturbances, of the earth's magnetic field and the existence of aurora and sunspots, the potential gradients become so great that grounded wire lines are seriously affected. The potentials set up add or subtract from the usual line battery and as these voltages on long lines may in severe cases be 400-500 volts or more, there is not only the problem of maintaining service, but also that of insuring adequate apparatus and cable protection and eliminating fire hazard due to insulation breakdown.

For the most part, the voltages are in the order of 0 to 50 volts and may change from positive to negative within a few seconds. The duration may be from a few minutes to several days and the

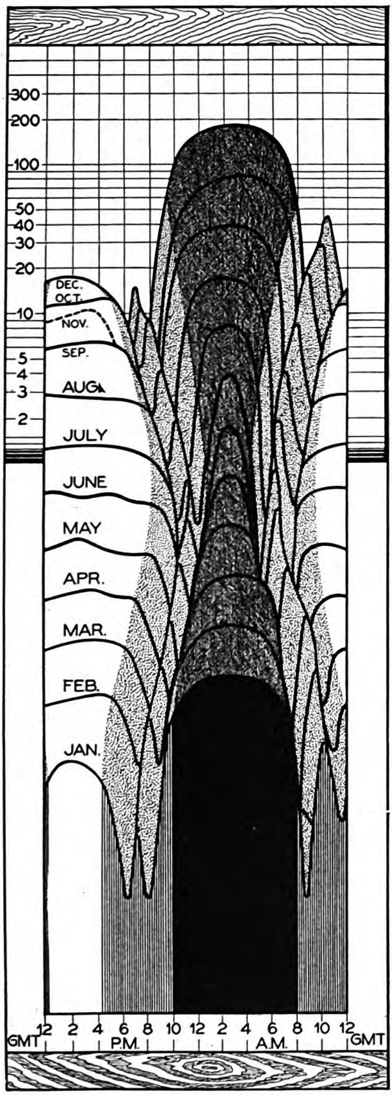


Fig. 26—Transatlantic Radio Telephone Transmission, Diurnal variation of assumed undisturbed radio field strength transmission. Marion, Mass. (WSO) to New Southgate, England. 5282 km., 25,700 cycles.

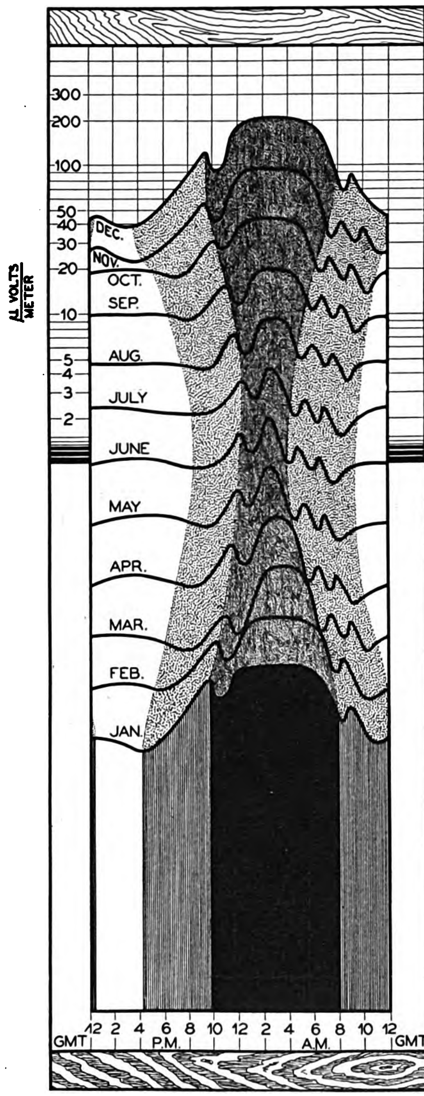


Fig. 27—Transatlantic Radio Telephone Transmission. Diurnal variation of assumed, undisturbed signal field strength transmission. Rocky Point, Long Island (WQL) to New Southgate, England. 5482 km., 17,130 cycles.

TABLE I

EARTH CURRENTS AND MAGNETIC AND SUNSPOT CONDITIONS

Date	Area Disturbed by Earth Currents	Severity	Time (EST) Beginning	Time (EST) End	Magnetic Character	Disturbance Beginning	Disturbance End (EST)	Provisional Sunspot Number
Apr. 28, 1921	Minneapolis Minn.	Mild	8:25 A.M.	10:00 A.M.	1.94	—	—	22
May 13–17, 1921	General—especially severe in New England	Periods of greatest intensity from 3 P.M. to 4 or 5 A.M. Disturbance most severe at about 9 P.M. May 14. Heat coils burned out, carbons grounded (operate on 350 volts), at Pittsburgh an arc of ¼ in. to 1 in. was noted between carbons. Condensers tested for 500 volts were broken down so that voltages in excess of this must have occurred. Cable insulation was broken down and a fire started in the Philadelphia office. Brilliant aurora on night of 14th.	8:50 A.M. (May 13)	12:30 A.M. (May 17)	20 Needle thrown out of adjustment so that complete record not available for May 13 to May 16. Character for May 12, 4.33. May 17, 3.32.	8:09 A.M. May 13	3:00 A.M. May 17	May 13–50 May 14–71 May 15–63 May 16–49 May 17–37
May 20, 1921	Pennsylvania to Missouri	Mild	12:55 P.M.	4:04 P.M.	5.50	9:33 A.M.	Midnight	18
Sept. 2, 1921	Pittsburgh, Pa.	Very mild	During night	to Sept. 2	3.99	11:00 A.M. Sept. 1	12:00 Noon Sept. 2	9
Feb. 13, 1922	Philadelphia to New York	Very mild	6:00 P.M.	8:00 P.M.	1.42	3:00 A.M. Feb. 14		63
Oct. 16, 1923	Indianapolis to Chicago	Mild	8:15 P.M.	9:40 P.M.	4.5	2:00 A.M. Oct. 15	2:00 A.M. Oct. 18	8
Oct. 27, 1923	Ohio	Mild	8:10 A.M.	9:14 A.M.	1.4	—	—	21
Jan. 29, 1924	N.E. quarter of Mississippi Valley	Quite severe with maximum disturbance the first hour.	1:00 P.M.	8:45 P.M.	5.8	12:24 A.M.	10:00 P.M.	0
May 9, 1925	Missouri	Mild	6:05 P.M.	9:25 P.M.	1.9	4:00 P.M. May 8	5:00 P.M. May 9	43
June 24, 1925	Wisconsin	Mild	3:18 P.M.	6:00 P.M.	4.55	8:00 P.M. June 22	6:00 P.M. June 25	7
Jan. 26, 1926	General — Except New England	Quite severe	11:19 A.M.	7:15 P.M.	9.1	11:17 A.M.	Midnight	105
Feb. 23, 1926	General — Except New England	Quite severe	9:05 A.M.	6:19 P.M.	5.1 (14 on Feb. 24)	11:26 A.M. Feb. 23	11:00 P.M. Feb. 24	48
Mar. 5, 1926	Middle West	Rather severe	9:25 A.M.	2:20 P.M.	7.55	5:03 A.M. Mar. 5	3:00 A.M. Mar. 6	119
Apr. 14, 1926	New England to Minnesota	Rather severe	9:45 A.M.	10:57 P.M.	13.65	9:00 A.M. Apr. 14	10:00 P.M. Apr. 17	71
Oct. 15, 1926	General	Very severe. At many observation points there was a lull between about 11:30 A.M. and 2:00 P.M. Aurora borealis display at night after disturbance had ceased.	8:45 A.M.	11:40 P.M.	20 (Estimated) Needle thrown out of adjustment	4:00 P.M. (Oct. 14) Most violent from noon to 4:00 P.M. Oct. 15	Midnight Oct. 15	112

P.M. hemisphere seems to be most affected. The disturbances may be comparatively local in character or may extend over the whole country with the effect about one-third as great on the

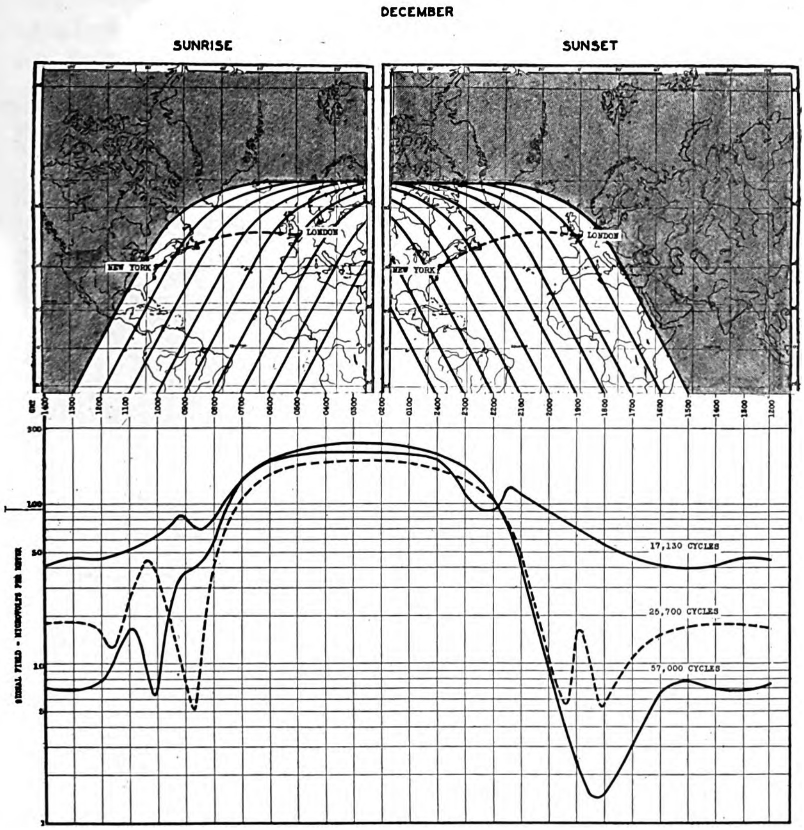


Fig. 28—Relation of Sunset and Sunrise to Transatlantic Radio Transmission, December.

east-west lines as on the north-south lines; the more severe the disturbance, usually the larger the area disturbed.

Table 1 gives the data on all disturbances which have been noted on the telegraph circuits of the American Telephone and Telegraph Company for the years 1921 to 1926, inclusive. The number of storms is, after all, comparatively small. The respective sunspot numbers range from 0 to 119, averaging possibly 50. The disturbance for the day with no sunspots was quite severe, and

on the other hand, no disturbances were noted on 300 days during 1923 to 1926 when sunspot numbers have exceeded 50, even up to 160.

Mild disturbances on telegraph lines have been noted with quiet magnetic conditions, but every severe disturbance has been

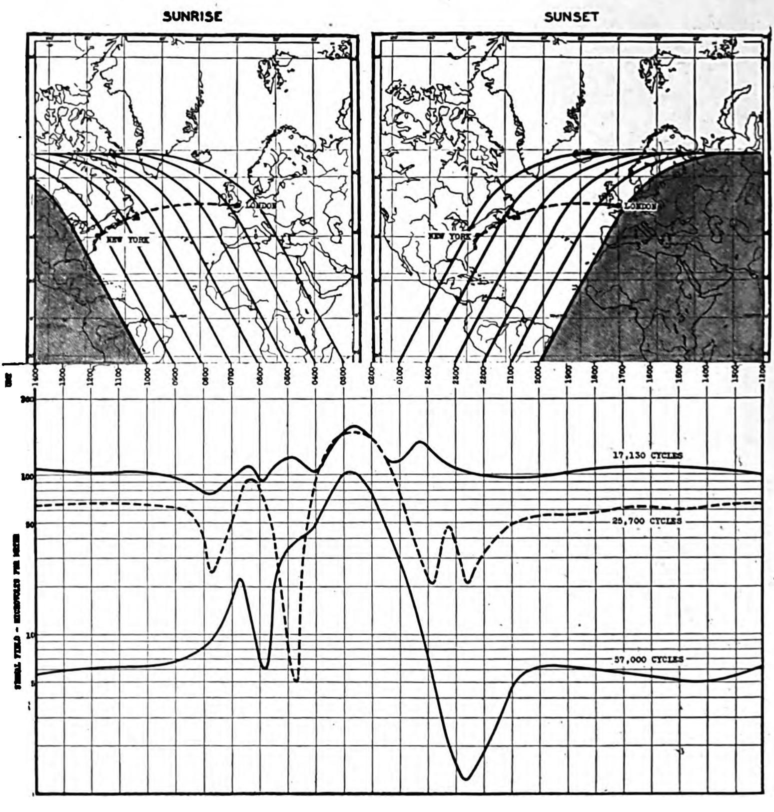


Fig. 29—Relation of Sunset and Sunrise to Transatlantic Radio Transmission, June.

accompanied by a severe magnetic disturbance. The extremely severe magnetic disturbances are also accompanied by earth current disturbances, but this does not hold for the less severe. For example, the magnetic character for January 29, 1924, with quite a severe earth current disturbance was 5.8. During the four years 1923-1926, 18 out of 23 days with magnetic character above 5.8 (some as high as 10) were unaccompanied by wire line distur-

bance. In most cases line disturbances occur during the period between the beginning and end of the magnetic disturbances. However, on February 13, 1922, the trouble on the wire circuits occurred between 6 P.M., and 8 P.M., while the magnetic disturbance, which was slight, did not begin until 3 A.M. the next morning. On January 23, 1926, the wire line disturbance began over two hours before the magnetic disturbance and was essentially over by 6

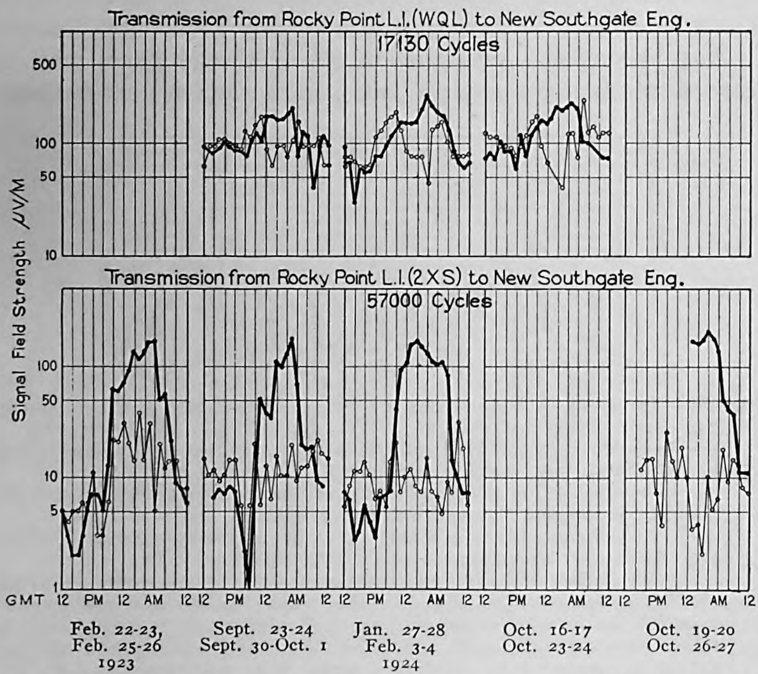


Fig. 30—Transatlantic Radio Telephone Measurements; Effect of Solar Disturbances on Radio Transmission.
——— More or less normal transmission week-end before disturbances occurred.
——— Abnormal transmission on following week-end, disturbance still in progress.

P.M., while the most violent disturbance in the earth's field occurred the next day, February 24, 1926. Another point of interest is that on October 15, 1926, the lull in the wire line disturbance from 11:30 A.M. to 2:30 P.M. occurred during the period of most violent disturbance of the earth's field-noon to 4:00 P.M.

The results seem to bear out the conclusions reached by the investigators of the Carnegie Institution.

Radio Transmission

General. In connection with the experiments prior to the establishment of a commercial radio-telephone circuit, considerable amount of data was accumulated on radio transmission across the Atlantic. Most of these data were on a frequency in the neighborhood of 60 kilocycles (5,000 meters) although some information was obtained at frequencies in the neighborhood of 25 kilocycles (12,500 meters) and 17 kilocycles (17,500 meters). The greater part of these data has already been published (9).

For convenience, transmission during the 24 hours may be divided into four periods: viz., the period when the entire trans-

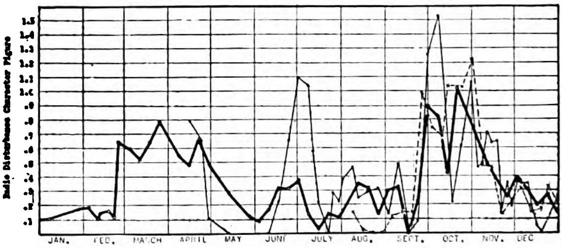


Fig. 31—Variation in Disturbance of Night Time Radio Field Strength, 1923.

———— Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km., 300 amperes. Antenna current, 57 kc.

———— Rocky Point, L.I. (WQL) to New Southgate, England, 5480 km., 600 amperes. Antenna current, 17.1 kc.

- - - - - - - Marion, Mass. (WSO) to New Southgate, England, 5280 km., 600 amperes. Antenna current, 25.7 kc.

These radio character figures are a measure of the extent to which night time signal fields are decreased from the assumed undisturbed value to the daylight value. If the radio character is zero, the night time fields are normal; if unity, the night fields are equal to the day fields; if greater than unity, the night fields are even less than the day fields.

mission path (great circle) is in daylight, the period when the entire path is in darkness, and the two transition periods, one in the morning and one at night. Fig. 18 shows the diurnal variations in signal field strength for the various months of the year for transmission from Rocky Point, Long Island, to New Southgate, England, on a frequency of 57,000 cycles. In order to show all the essentials, the figure lost its guide of perspective. One must visualize, therefore, a three-dimensional figure in which the vertical plane showing the field-strength grid is at the back. The months of the year are distributed along the base and the third

dimension is the time of day. The shaded portions indicate the extent to which the transmission path is in darkness.

The night period is one of relatively high, and usually unstable, field-strength values. They are often thought of as representing the unabsorbed transmission and expressed by the Hertzian expansion formula $E(\mu v/M)=\frac{120\pi HI}{\lambda D}$ commonly spoken of as the Inverse-Distance Law.

The daylight period is one of relatively low, although as a rule quite stable, field-strength values. These are the values to which

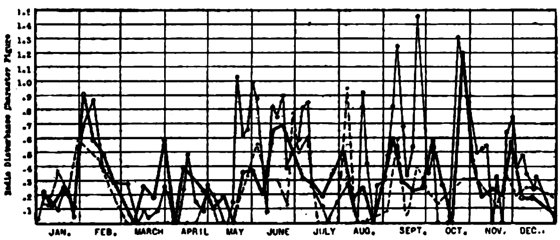


Fig. 32—Variation in Disturbance of Night Time Radio Field Strength, 1924.

————Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km., 300 amperes. Antenna current, 57 kc.

————Rocky Point, L.I. (WQL) to New Southgate, England, 5480 km., 600 amperes. Antenna current, 17.1 kc.

------- Marion Mass. (WSO) to New Southgate, England, 5280 km., 600 amperes. Antenna current, 25.7 kc.

These radio character figures are a measure of the extent to which night time signal fields are decreased from the assumed undisturbed value to the daylight value. If the radio character is zero, the night time fields are normal; if unity, the night time fields are equal to the day fields; if greater than unity, the night fields are even less than the day fields.

various transmission formulas apply. The more widely used of these formulas usually consist of the expression for the Inverse-Distance Law and an exponential factor representing the absorption encountered during the day.

During the transition periods, the change from daylight field to night field or vice versa is not a direct one, but passes through one or more periods of equivalent high attenuation.

Night-Time Radio Transmission. For the purpose of comparing radio transmission from day to day, the first requisite was to determine the standard for comparison. It has already been noted, during periods of solar disturbances, that the night fields were

decreased greatly and that the daylight fields were somewhat increased. The first step, then, in attempting to generalize our data, was to obtain averages of 24-hour transmission for all days which might be characterized as not seriously disturbed. This was done for each individual month. It was next found that, for the night-time fields, smoother curves and more consistent results were obtained by plotting, for each month, the maximum values ever obtained for each particular hour. It is of interest to note in this connection that for frequencies of 17,000 cycles and 25,000 cycles as high night-time values were obtained during the summer months as during the winter months. This maximum value was

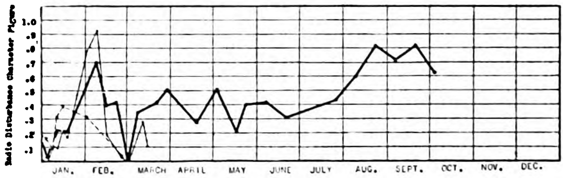


Fig. 33—Variation in Disturbance of Night Time Radio Field Strength, 1925.

——— Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km., 300 amperes. Antenna current, 57 kc.

——— Rocky Point, L.I. (WQL) to New Southgate, England, 5480 km., 600 amperes. Antenna current, 17.1 kc.

- - - - - - - Marion, Mass. (WSO) to New Southgate, England, 5280 km., 600 amperes. Antenna current, 25.7 kc.

These radio character figures are a measure of the extent to which night time signal fields are decreased from the assumed undisturbed value to the daylight value. If the radio character is zero, the night time fields are normal; if unity, the night fields are equal to the day fields; if greater than unity, the night fields are even less than the day fields.

essentially that calculated by the Inverse-Distance formula. On 57,000 cycles, values as high as those calculated by that formula are obtained only during the winter months.

Daylight Radio Transmission. The above method appears to eliminate successfully the effect of disturbed transmission for the night period. For the daylight period, the average field for each hour was taken as computed from the "quiet" days referred to above. This minimizes, but, as seen later, does not entirely eliminate the effect of solar disturbances. As these daylight values are used only for computing the amount of night-time disturbance, and the difference between the "quiet" and "affected" daylight values is not a particularly large one, it was thought that the day-

light values as obtained above, would represent conditions for the present at least, as well as any obtained by further assumptions.

As is well known, the daylight fields for equal radiated powers are higher, the lower the frequency. This relationship, based on measurements taken on three frequencies in the particular frequency range of 60 to 15 kilocycles and over the transmission path New York and London can be expressed by the empirical formula

$$E_{\mu v}/M = \frac{120\pi HI}{\lambda D} e^{\frac{-0.005D}{\lambda^{1.25}}}$$

$$= \sqrt{\frac{P_{kw} 3.10^5}{D}}\, e^{\frac{-0.005D}{\lambda^{1.25}}}$$

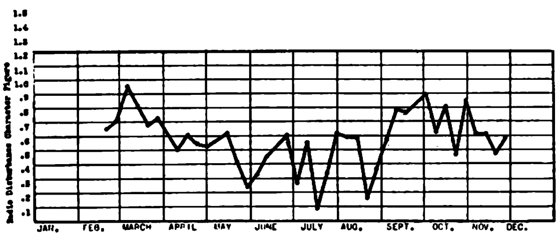


Fig. 34—Variation in Disturbance of Night Time Radio Field Strength, 1926. Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km., 300 amperes. Antenna current, 57 kc. These radio character figures are a measure of the extent to which night time signal fields are decreased from the assumed undisturbed value to the daylight value. If the radio character is zero, the night time fields are normal; if unity, the night fields are equal to the day fields; if greater than unity, the night fields are even less than the day fields.

This formula has also been found to represent quite well transmission data obtained by engineers of the Marconi Company on a trip around the world in 1922 and 1923 (13). One apparent exception, however, to increasing fields with decreasing frequency seems to be indicated by data obtained several years ago in which abnormally low field strengths were obtained on frequencies in the vicinity of 40 kilocycles (Fig. 19). The study was made during the summer months and was suggested by exceptionally low fields experienced during the spring months on a frequency of 43 kilocycles. The field measurements at 58.5 kilocycles were taken at a period during which the earth's field was disturbed and hence are somewhat higher than average.

Curves showing the variation in daylight field for such measurements as were made in 1923 to 1926, inclusive, on 60,000 cycles, 25,000 cycles and 17,000 cycles are shown in Figs. 20 to 23, inclusive.

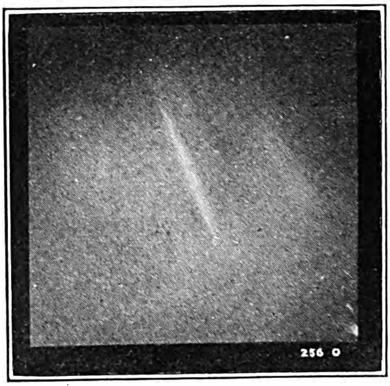


Fig. 35—Photograph of Aurora at Oslo, Norway, May 12, 1921 (Stormer).

Sunrise and Sunset Effects on Radio Transmission. Some difficulty was encountered in obtaining a picture of the variation in field during the transition period due to the fact that a large part of the phenomena is of short duration as compared with the

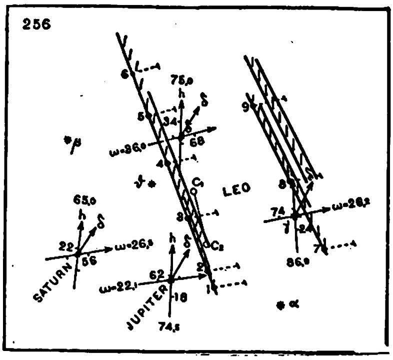


Fig. 36—Sketch of Aurora (Fig. 35) with Background of Stars.

interval between measurements. Accordingly, the data obtained were plotted with respect to sunrise and sunset instead of standard time.

Fig. 24 shows the results obtained for the transition period as daylight recedes from the transmission path. The assumed un-

disturbed variations for the three frequencies 17,130 cycles, 25,700 cycles and 57,000 cycles are shown by the respective solid lines. The major sunset dips occur approximately 2 hours and 30 minutes after sunset in London for the frequencies 17,130 cycles and 57,000 cycles and the two dips occur at about 2 hours and 15 minutes and 3 hours and 50 minutes after sunset in London for 25,700

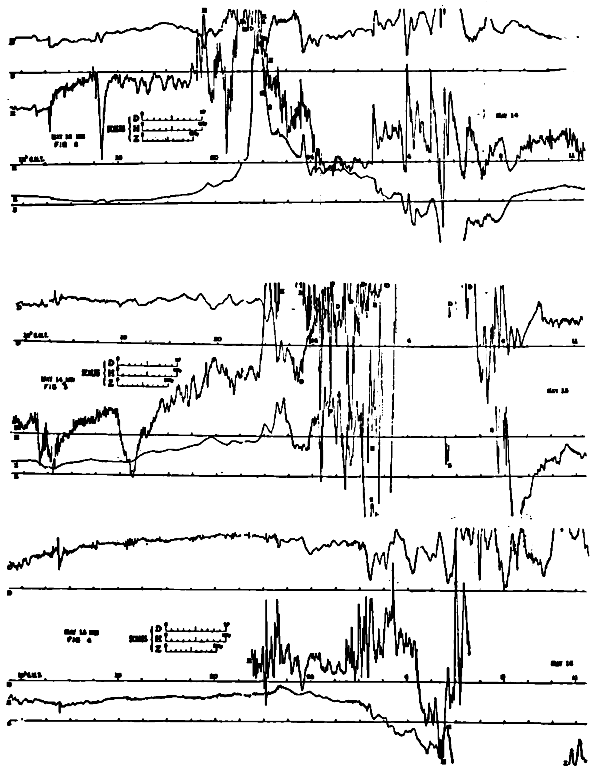


Fig. 37—Magnetograms of Earth's Magnetic Field, Cheltenham, Maryland. (From U. S. Coast and Geodetic Survey Report Serial 275).

cycles, irrespective of seasons. These are independent of season because the portion of great circle path between New York and London, which is being plunged into darkness during the first three or four hours after sunset in England, is at approximately the same latitude as London. In winter, sunset at New York occurs 5 hours and 40 minutes after sunset at London and in summer only 4 hours and 15 minutes after sunset at London. The

phenomena associated with sunset at New York can thus be segregated. One of them is the dip in transmission (17 kilocycles)

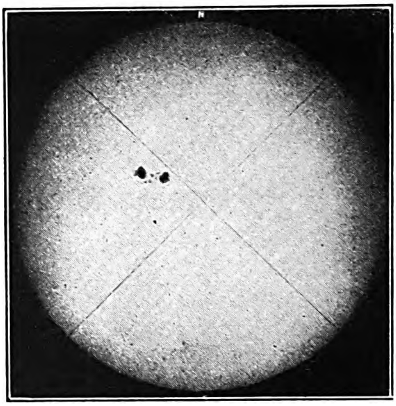


Fig. 38—Photograph of Solar Disk, May 13, 1921.

occurring approximately 30 to 50 minutes after sunset at New York.

The dotted curves indicate the transmission tendency during disturbed periods. For severe conditions this tendency is for the

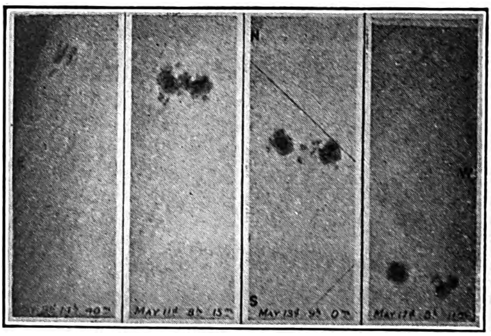


Fig. 39—Photograph of Solar Disk showing Progress of Sunspot Group.

sunset dip to be less pronounced and for the field strengths to drop to the low night values directly after sunset in New York.

Fig. 25 shows similar results obtained for the sunrise transition period. The same tendency to smooth out the peaks and dips is shown. In one case of severe disturbance, a complete reversal was found.

As many of the phenomena accompanying sunrise and sunset are of much shorter duration than those which the transmission schedules were designed to study, there are, no doubt, other variations which our data can only suggest. For example, one piece of evidence at hand shows a decrease in the field (57,000 cycles) by the factor of 3 over a period of only 5 to 10 minutes at sunset at New York. For the same reason, the magnitude of the dips is no doubt underestimated. One continuous record indicates a fall in field strength (57,000 cycles) to half value at sunrise England whereas our other data only warranted showing a small depression.

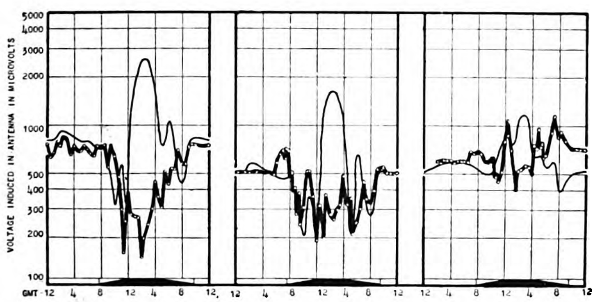


Fig. 40—Comparison of Radio Transmission During Disturbed Period in May, 1921 and Assumed Undisturbed Transmission Data at Chelmsford, England. Taken by Engineers of Marconi Wireless Telegraph Company, Ltd.

New Brunswick, N. J. to Chelmsford, England, 13.6 km. Distance 5420 km. May 13 to May 14, 1921. Light curve indicates assumed undisturbed transmission from Marion, Mass. to London, England, for May. Approximately same wavelength and distance.

Marion, Mass. to Chelmsford, England 11.6 km. Distance 5310 km. May 19 to May 20, 1921. Light line indicates assumed undisturbed transmission from Marion, Mass. to London, England, for May. Same wavelength and approximately same distance.

Tuckerton, N. J. to Chelmsford, England, 16 km. Distance 5520 km. May 23 to May 24, 1921. Light line indicates assumed undisturbed transmission from Rocky Point, L. I. to London, England, for May. Approximately same wavelength and distance.

Because radio transmission is so sensitive to light conditions during these transition periods, a study of this phase of radio transmission may prove to be quite profitable.

Generalized Diurnal Variation of Undisturbed Radio Transmission. With the night fields determined by the envelope of maximum values for each month, the day fields obtained by averages, and the major variations during the transition sunset and sunrise

periods at least qualitatively known, composite curves of assumed undisturbed transmission could be constructed for each month of the year as shown in Fig. 18. Similar figures for transmission on 25,700 cycles and 17,130 cycles are shown in Figs. 26 and 27.

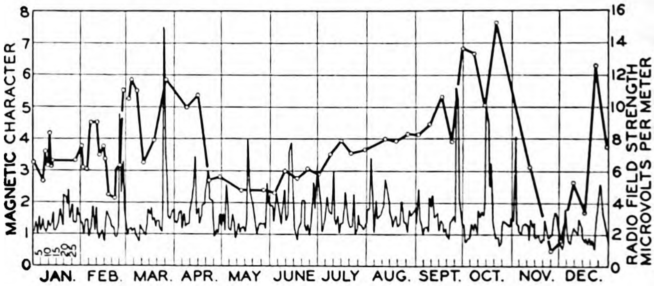


Fig. 41—Comparison of Average Daylight Field Strength with Character Figure of Earth's Magnetic Field.
———Radio transmission from Rocky Point, L.I. (57 kc.) to New Southgate, England.
———Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1923.

Fig. 28 shows the positions of the sunset and sunrise shadow walls for December in relation to the transmission path and the corresponding variation in the radio field while Fig. 29 shows similarly the conditions for June.

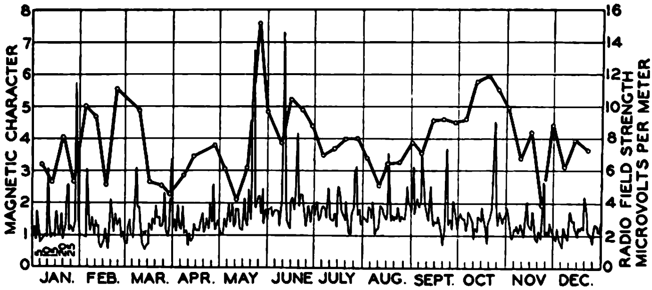


Fig. 42—Comparison of Average Daylight Field Strength with Character Figure of Earth's Magnetic Field.
———Radio transmission from Rocky Point, L.I. (57 kc.) to New Southgate, England.
———Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1924.

Measure of Night-Time Radio Disturbance. As shown in an earlier paper (9), there occur periods during which the night fields are very much reduced and subject to considerable fluctua-

tion and change of direction. These periods were shown to coincide with periods of marked disturbance in the earth's magnetic field. Fig. 30 is reproduced from the earlier paper referred to.

In order to obtain radio transmission character figures which would be more or less independent of the frequency measured, the method suggesting itself was that of using the extent to which the night values were reduced to the normal daylight values. In order to minimize as much as possible the effect of fluctuations which would tend to make the measurements non-representative, the character figure finally decided upon was the ratio of which the area (semi-log coordinates) of the curve enclosed by the assumed undisturbed night values less the average daylight value

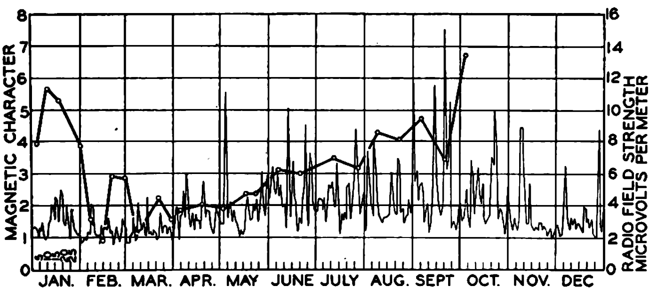


Fig. 43—Comparison of Average Daylight Field Strength with Character Figure of Earth's Magnetic Field.
———Radio transmission from Rocky Point, L.I. (57 kc.) to New Southgate, England.
———Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1925.

is the denominator and the area represented by the difference between the area of assumed undisturbed night values and the observed night values received is the numerator. The actual method of arriving at this figure was to obtain the sum of the logarithms of all the night ordinates of the disturbed N and the assumed undisturbed N curves and the log of the average daylight field d.

The radio character figure is then equal to $\dfrac{\log N - \log N'}{\log N - n \log d}$ where n is the number of ordinates. If the night fields are equal to those assumed undisturbed transmission, the character figure is zero; if night fields are reduced to the day fields, the character figure is unity; if reduced below the daylight value, the character figure

is greater than unity. Because of the short period during the summer months in which the entire transatlantic path is in darkness, only 3 or 4 night-time values of field strength represent transmission conditions. As these fall on the side of the curve where field strengths are rapidly changing, considerable error may be introduced. A single experimental error or a measurement made at an unusually favorable or unfavorable moment might change altogether the radio character figure for that day.

Curves showing the variation of character figures for such measurements as were taken from 1923 to 1926, inclusive, on

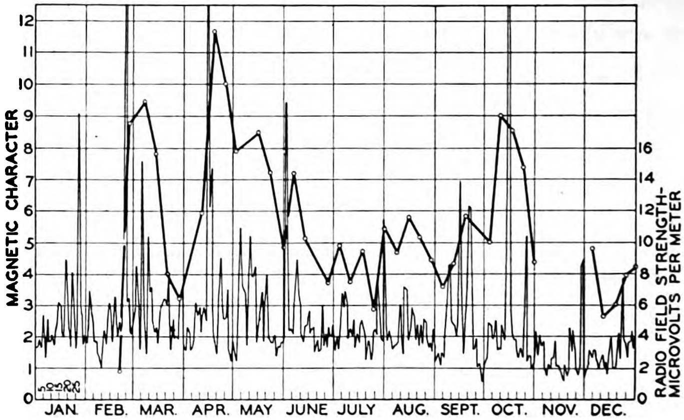


Fig. 44—Comparison of Average Daylight Field Strength with Character Figure of Earth's Magnetic Field.
————Radio transmission from Rocky Point, L.I. (57 kc.) to Wroughton, England.
————Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1926.

60,000 cycles, 25,000 cycles and 17,000 cycles over the transmission path between roughly New York and London are shown in Figs. 31, 32, 33, and 34.

III. Intercorrelation

Disturbance of May, 1921. One of the most severe disturbances of recent years occurred May 13–17, 1921. Auroras were particularly brilliant and observed at very low latitudes. One of the numerous photographs obtained in Norway (8) is shown in Fig. 35 and a sketch of its position against a background of stars is shown in Fig. 36. The calculated position is shown on the map (S-260)

of Fig. 16 with a height extending from 103 to 460 kilometers. During the storm, grounded telegraph circuits were rendered inoperative due to excessive earth potentials which burned out line protectors, broke down condensers and insulation and even started fires in terminal offices.

The disturbance in the earth's field was also violent, with the magnetometer trace off scale for hours at a time at observatories in both England and America. At Cheltenham, Maryland, the needle was twice thrown out of balance. The magnetograms (2) at Cheltenham for May 13–16 are shown in Fig. 37 and indicate the nature of such disturbances.

Fig. 38 shows a picture of the solar disk taken at the Greenwich Observatory, London, at 0900 GMT on the day the storm began

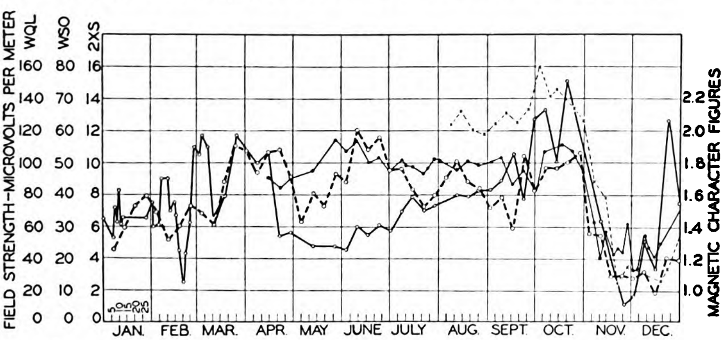


Fig. 45—Comparison of Average Daylight Field Strength and Weekly 3-Week Moving Averages of Magnetic Character Figures, 1923.
——— Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km.—300 amperes. Antenna current, 57 kc.
——— Rocky Point, L.I. (WQL) to New Southgate, England, 5480 km.—600 amperes. Antenna current, 17.1 kc.
- - - - - - Marion, Mass. (WSO) to New Southgate, England, 5280 km. —600 amperes. Antenna current, 25.7 kc.

(12). It is to be noted that the end of the sun's axis is 21 deg. west of north and the sun's equator $2\frac{1}{2}$ deg. north of the center of the disk. The sunspot group is on the sun's equator and the largest in this position, in the last half-century. The area is 1/1500 of the sun's surface or eight times the area of the earth.

The spot was nearest the center of the disk May 14 at 1600 GMT when it was within 3 deg. It was then nearly in line with the earth, but the magnetic storm occurred 27 hours earlier. At this time the leading spot was $11\frac{1}{2}$ deg. east of the central meridian and the following spot 19 deg. The greatest intensity

of the disturbance was about 0500 on May 15, at which time the following spot was 3 deg. past the central meridian. The second maximum occurred at 0800 GMT, May 16.

Fig. 39 shows the progress of the group across the sun's disk from the day when it first appeared—May 8.

Fortunately data on radio transmission for this period are also available. Fig. 40 shows variation in signal field from New Brunswick, N. J., Marion, Mass., and Tuckerton, N. J., as measured by the engineers of the Marconi Wireless Telegraph Company, Ltd., at Chelmsford, England, (13). Curves of assumed undisturbed radio transmission from Marion and Rocky Point of approximately the same frequencies and distances are also

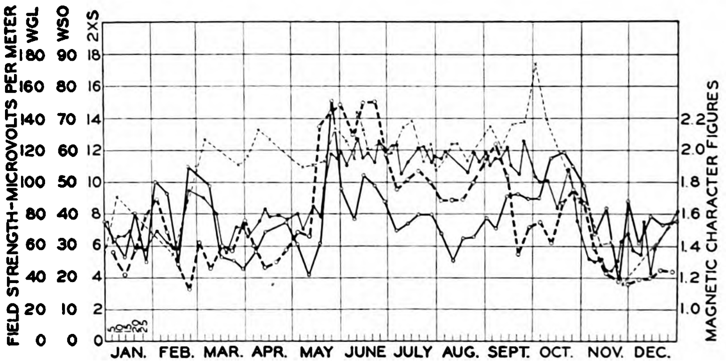


Fig. 46—Comparison of Average Daylight Field Strength and Weekly 3-Week Moving Averages of Magnetic Character Figures, 1924.
————Rocky Point, L.I. (2XS) to New Southgate, England, 5480 km.—300 amperes. Antenna current, 57 kc.
————Rocky Point, L.I. (WQL) to New Southgate, England, 5480 km.—600 amperes. Antenna current, 17.1 kc.
- - - - - - -Marion, Mass. (WSO) to New Southgate, England, 5280 km.—600 amperes. Antenna current, 25.7 kc.

drawn in for comparison. It is seen that the night-time transmission is reduced to one-half and one-third the values which might be expected.

Earth Currents. In the case just cited auroras, earth currents, sun spots, earth's field, and radio were all severely affected. The agreement is not, however, always so good. As previously pointed out in connection with earth currents on telegraph lines, severe disturbances may occur with no visible sun spots. On the other hand no disturbance occurred on 300 days in 1923 to 1926 on which the sunspot numbers exceeded the 50 which obtained on

May 13, when the most severe disturbance on record occurred on the telegraph lines. Discrepancies in the times of beginning and in times of maximum disturbances in earth currents and earth's magnetic field support conclusions reached by other investigators that these phenomena are not intimately connected.

Sun Spots and Magnetic Disturbances. Comparison of sunspot numbers (Figs. 2 to 5) and magnetic character figures (Figs. 10 to 13) for 1923 and 1926, inclusive, shows that the severe disturbances of February 25, 1923, and January 29, 1924, were unaccompanied by any sun spots. The severe disturbances of February 24, and April 15, 1926, as well as all others in the previous years (1923–1926) and including the severe disturbance of May, 1921, were accompanied by relatively low sunspot numbers as prevailed

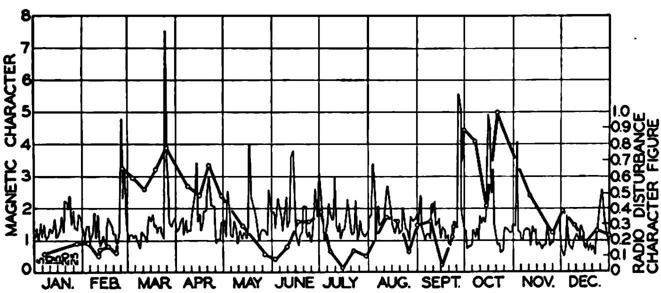


Fig. 47—Comparison of Disturbance in Night Time Radio Transmission with Character Figure of Earth's Magnetic Field.
———— Radio Transmission from Rocky Point, L.I. (57 kc.) to New Southgate, England.
———— Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1923.

at other times in 1925 and 1926 when no severe magnetic disturbances occurred. It is true that these discrepancies might be explained by the fact that at times the sun spots were unfavorably situated with respect to the earth, i.e., not in or near that portion of the sun's surface cut by a cone of which the center of the sun is the apex and the earth's disk the base.

Daylight Radio Transmission and Magnetic Disturbances. Figs. 41 to 44 show the comparison of the average daylight radio field strength (57,000 cycles) with daily magnetic character figures. It is seen that the high radio fields do not occur particularly on days of high magnetic character but rather during periods when magnetic storms occur. Because the radio data are available for one day a week only, no detailed conclusions can be drawn. For

the most part, however, the high fields follow the magnetic disturbance and then gradually fall off, although the gradual rise in field strengths during August and September in 1923 and during October in 1924 seems to be independent of magnetic disturbances. One outstanding case of the high field strength preceding the magnetic disturbances is that of October 10, 1926, where the magnetic disturbance did not commence until the third day after and did not reach its maximum until the fifth day after.

A comparison of radio transmission with three-week moving averages as in Figs. 45 and 46 does not show any great degree of correlation except in the major variation. The agreement with radio fields on the various frequencies is, however, about as good

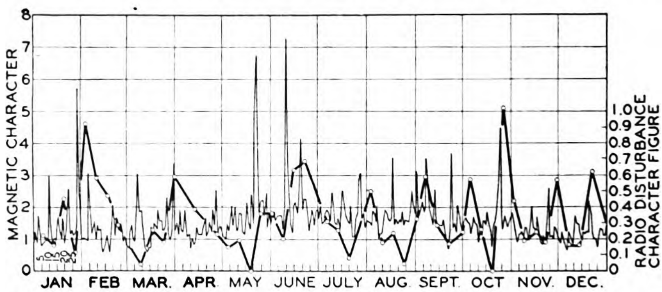


Fig. 48—Comparison of Disturbance in Night Time Radio Transmission with Character Figure of Earth's Magnetic Field.
———— Radio Transmission from Rocky Point, L.I. (57 kc.) to New Southgate, England.
———— Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1924.

as between transmission on various frequencies. One outstanding feature which is shared by transmission on all frequencies and magnetic characters alike is the low fields and low disturbance for the two to three weeks on either side of December 1. This may be of significance in view of the sun's equator (region of low activity) being on the line of centers of the earth and sun on December 7.

Night Radio Transmission and Magnetic Disturbances. The comparison of disturbances of night-time radio transmission and earth's magnetic field is shown in Figs. 47 to 50. As before, daily comparisons tell us little. There seems to be a close relation between the radio disturbances and the magnetic disturbances, however, in that the radio disturbances are consistently low preceding a magnetic disturbance, then there is an abrupt increase

accompanying the magnetic disturbance and finally a gradual recuperation which may last several months depending on the severity. This "hang-over" effect is seen best in the figures for 1923 and 1924 where disturbances are less frequent, permitting recovery to take place. Disturbances occurring before recovery has taken place superimpose their effects upon conditions present at the time.

Because of this "hangover" effect it is not entirely correct to compare magnitudes of radio disturbances with magnitudes of magnetic disturbances as the latter might occur partly in one day and partly in the next so that although the magnetic character of both days may be relatively low, the magnetic character of the 24 hours including the period of maximum might be considerably

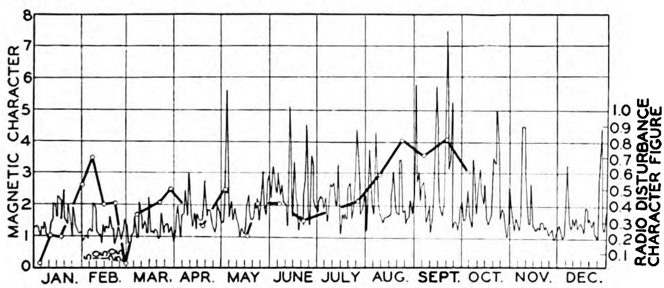


Fig. 49—Comparison of Disturbance in Night Time Radio Transmission with Character Figure of Earth's Magnetic Field.
——— Radio Transmission from Rocky Point, L.I. (57 kc.) to New Southgate, England.
——— Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1925.

higher. Such is the case of September 26 and 27, 1923, with character figures of 5.6 and 5.2 respectively. Taking the 24 hours of maximum activity which in this case was from noon on September 26, to noon on September 27, the magnetic character is 9.3.

A better picture of the effect of a disturbance and the recovery is shown in Fig. 51. Each curve represents the diurnal variation of signal field for the 24 hours commencing at noon one day and ending at noon the next. Radio transmission on January 27–28 was quite normal with a radio character figure of 0.1. The disturbance in the earth's field occurred January 29, after which the first radio test schedule on February 3–4 showed a radio disturbance character figure of 0.92. The subsequent tests show the gradual recovery with character figures of 0.58, 0.47, 0.28, and 0.04. This process took approximately 5 weeks.

On the lower frequencies, the recovery appears to be more rapid and there is some evidence that there is a lag in the maximum radio disturbance behind the disturbance of the earth's field.

The disturbances of night-time radio fields can, therefore, be reasonably attributed to some cause which simultaneously affects the earth's field; but the rates of recovery in the two cases are different, being much more rapid in the case of the earth's field. Although the daylight radio fields are highest at periods of magnetic disturbance, they often increase toward these high values ahead of the magnetic and have been known to jump to

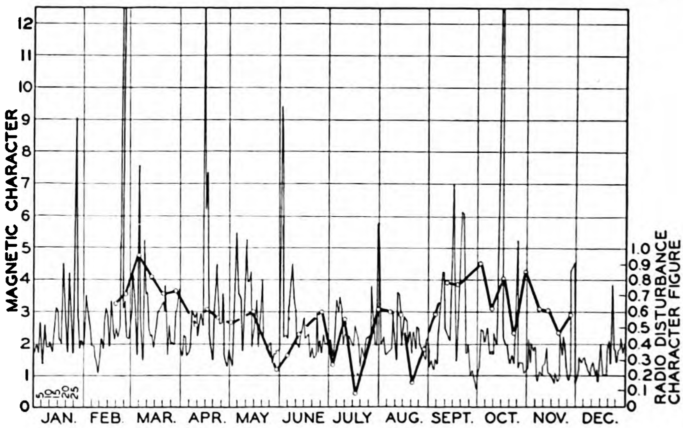


Fig. 50—Comparison of Disturbance in Night Time Radio Transmission with Character Figure of Earth's Magnetic Field.
————Radio Transmission from Rocky Point, L.I. (57 kc.) to Wroughton, England.
————Magnetic data taken at Cheltenham, Maryland. U. S. Coast and Geodetic Survey, 1926.
(The magnetic characters as represented above are determined by adding the total variation between hourly averages of horizontal and vertical field (gammas) for each day, and dividing by 100.)

the peak value several days before the magnetic disturbance. This would seem to indicate that the mechanism of the increased day fields and that of the decreased night values were different. The former might be affected by, say, corpuscular radiation but the latter by wave radiation.

It is seen that if for no other reason that than the rates of recovery are different, the mathematical correlation of day-to-day magnetic and radio conditions would probably not be high.

Fig. 52 shows three-week moving averages computed for every week for radio character (night), daylight radio fields, magnetic character solar constant and sun spot.

The dotted lines in each case represent the trend as computed from 18-month moving averages computed for every 6 months. The agreement of this trend is exceptionally good in all cases expect in the case of solar constant which shows a decrease in 1926 while the other factors maintain their maximum. As the data only extend over a part of the cycle of solar activity, no definite conclusions can be drawn as to whether this apparent

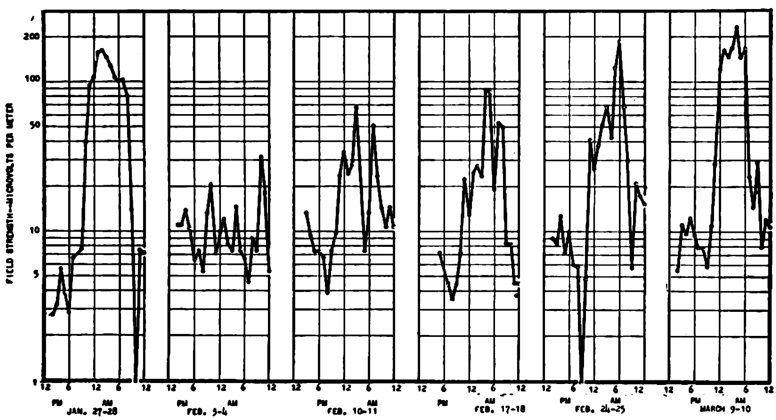


Fig. 51—Transatlantic Radio Telephone Transmission. Diurnal variation in signal field on successive week-ends showing recovery after disturbance January 29, 1924. Rocky Point, L. I. to New Southgate, England. 5480 km., 57,000 cycles.

phase displacement is real or not. The results of detailed mathematical correlation can best be summarized by Table II.

The Correlation Coefficients are computed by the formula developed by Pearson

$$r = \frac{\Sigma(d_1 d_2)}{\sqrt{\Sigma d_1^2 \cdot \Sigma d_2^2}}$$

where d_1 and d_2 are the deviations of the individual values of two sets of data from their respective means. They endeavor to assign numerical values to the correlation which the eye attempts to do by inspection of the plotted curves. If the coefficient is unity and positive, the correlation is perfect and an increase in one is accompanied by an increase in the other. If the coefficient is

TABLE II

Intercorrelation of Various Factors Affected by Solar Activity

	Trend 6 Monthly 12-Month *Moving Averages*	4-Weekly 12-Week *Moving Averages*	4 Weekly 12-Week Moving Averages *(Trend Out)*	Weekly 3-Week Moving Averages *(Trend Out)*	Weekly Averages *(Trend Out)*
RADIO CHARACTER—NIGHT					
Magnetic Character	*+0.995* (406)	*+0.725* (15.4)	*+0.28* (2.8)	*+0.39* (9)	*+0.32* (6.5)
Sun Spots	*+0.985* (158)	*+0.664* (12.5)	*+0.29* (3.3)	*+0.22* (4.6)	*+0.15* (2.7)
Radio Fields—Day	*+0.975* (86.2)	*+0.633* (10.6)	*+0.55* (7.7)	*+0.37* (8.4)	*+0.34* (6.8)
Solar Constant	*+0.44* (2.5)	*+0.22* (2.9)	*−0.04* (0.4)	*+0.03* (0.6)	*−0.01* (0.2)
RADIO FIELDS—DAY					
Magnetic Character	*+0.987* (180)	*+0.695* (13.4)	*+0.38* (4.3)	*+0.41* (10.5)	*+0.33* (6.9)
Radio Character—Night	*+0.975* (86.2)	*+0.633* (10.6)	*+0.55* (7.7)	*+0.37* (8.4)	*+0.34* (6.8)
Sun Spots	*+0.95* (46.8)	*+0.54* (7 .7)	*+0.14* (1.4)	*+0.16* (3.4)	*+0.05* (0.9)
Solar Constant	*+0.30* (1.5)	*−0.07* (0.8)	*−0.09* (0.9)	*+0.03* (0.6)	*+0.06* (1.1)
MAGNETIC CHARACTER					
Radio Character—Night	*+0.995* (406)	*+0.725* (15.4)	*+0.28* (2.8)	*+0.38* (9)	*+0.32* (6.5)
Radio Fields—Day	*+0.987* (180)	*+0.695* (13.4)	*+0.38* (4.3)	*+0.41* (10.5)	*+0.33* (6.9)
Sun Spots	*+0.98* (127)	*+0.798* (22.8)	*+0.21* (2.3)	*+0.21* (4.6)	*+0.20* (4.3)
Solar Constant	*+0.39* (2.1)	*+0.14* (1.4)	*−0.46* (6)	*−0.21* (4.7)	*−0.14* (1.4
SUNSPOT NUMBERS					
Radio Character—Night	*+0.985* (158)	*+0.664* (12.5)	*+0.29* (3.3)	*+0.22* (4.6)	*+0.15* (2.7)
Magnetic Character	*+0.98* (127)	*+0.798* (22.8)	*+0.21* (2.3)	*+0.21* (4.6)	*+0.20* (4.3)
Radio Fields—Day	*+0.95* (46.8)	*+0.54* (7.7)	*+0.14* (1.4)	*+0.16* (3.4)	*+0.05* (0.9)
Solar Constant	*+0.53* (3.3)	*+0.39* (4.9)	*+0.1* (1.1)	*+0.23* (5.2)	*+0.11* (2.4)
SOLAR CONSTANT					
Sun Spots	*+0.53* (3.3)	*+0.39* (4.9)	*+0.1* (1.1)	*+0.23* (5.2)	*+0.11* (2.4)
Radio Character—Night	*+0.44* (2.5)	*+0.22* (2.9)	*−0.04* (0.4)	*+0.03* (0.6)	*−0.01* (0.2)
Magnetic Character	*+0.39* (2.1)	*+0.14* (1.4)	*−0.46* (6)	*−0.21* (4.7)	*−0.14* (1.4)
Radio Fields—Day	*+0.30* (1.5)	*−0.07* (0.8)	*−0.09* (0.9)	*+0.03* (0.6)	*+0.06* (1.1)

Italic figures denote Correlation Coefficient.
Figures in parentheses denote ratio of Correlation Coefficient to Probable Error.

unity and negative, the correlation is perfect but an increase in the one is accompanied by a decrease in the other. If the coefficient is zero, the variations in both are random and independent of each other. In order to determine the significance of the Correlation Coefficients, one resorts to the ratio of the coefficients to the

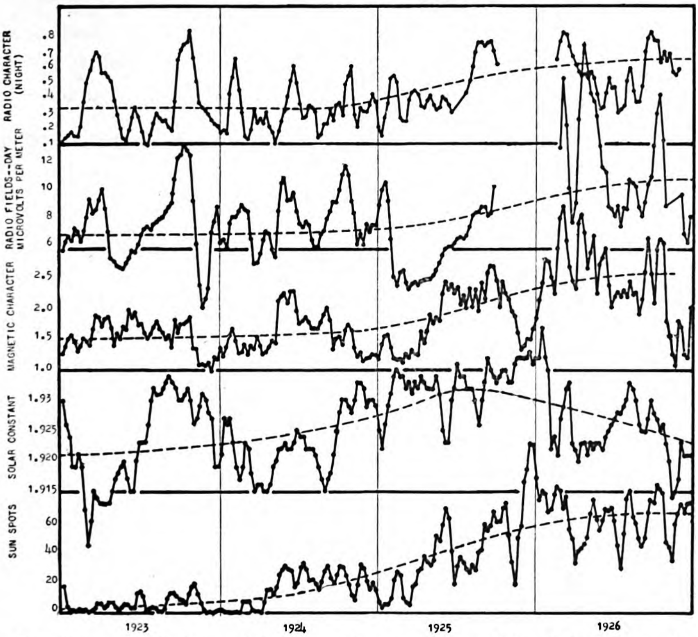


Fig. 52—Intercorrelation of Various Factors Affected by Solar Disturbances —Comparison of "Weekly" 3-Week Moving Averages.

Probable Error. The Probable Error is computed by the formula:

$$P.E = 0.6745\frac{1-r}{\sqrt{n}}$$

where r = correlation coefficient.
n = number of observations.

For practical purposes it is assumed that if the Correlation Coefficient is less than the Probable Error, there is no evidence of correlation; if the Correlation Coefficient is more than 6 times the Probable Error, correlation is good. In addition when the Probable Error is relatively small, there is decided correlation if the correlation coefficient is greater than 0.5 but not at all marked if less than 0.3.

It is seen from the table that whatever correlation exists lies in the broad movement corresponding to the eleven-year cycle of solar activity and that with day-to-day, week-to-week, or even three-month averages (Fig. 53) of deviations from this trend, the correlation is small. It is of interest to note that although the correlation of the trend of variation of the solar constant is positive, the correlation of some of the shorter period fluctuations is negative. This is especially evident in the case with magnetic activity.

In connection with correlation of radio transmission with solar

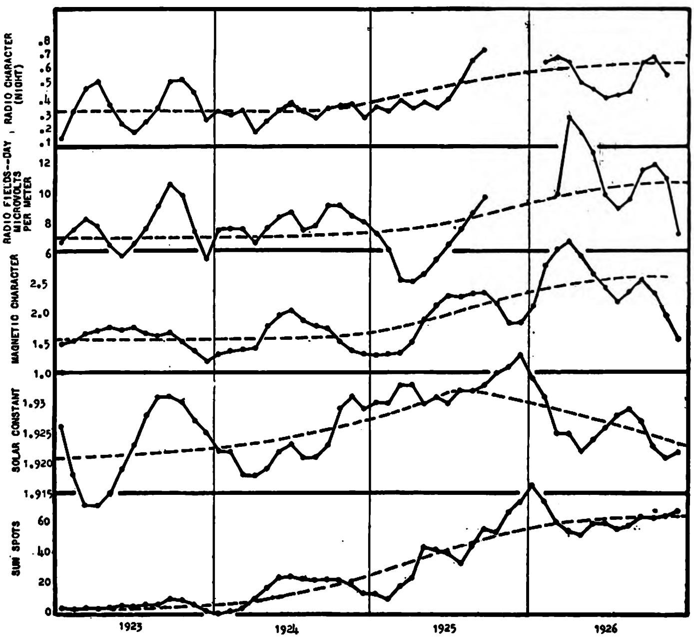


Fig. 53—Intercorrelation of Various Factors Affected by Solar Disturbances —Comparison of "Four Weekly" 12-Week Moving Averages.

activity for a complete cycle, there is only one set of consistent radio data in existence, that taken by Dr. Austin of the Bureau of Standards. Recently he has shown high correlation existing between the trends of radio transmission from Nauen for the years 1915 to 1926 inclusive and corresponding sunspot numbers.

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REPORT OF THE CHAIRMAN OF THE COMMISSION ON RADIO WAVE PROPAGATION,* INTERNATIONAL UNION OF SCIENTIFIC RADIO TELEGRAPHY

By

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THE work of the Commission on Radio Wave Propagation appointed at the last General Assembly of the International Union of Scientific Radio Telegraphy at Brussels in 1922, covers such a wide field of investigation that it seems best to confine this report to a discussion of the subjects on which the various workers are not yet entirely in accord, or those in which the conclusions are not yet definitely established.

A considerable portion of the work to be discussed has not been done directly by the Union and some of the investigations mentioned have been carried out in countries not connected with the organization.

Theory of Wave Propagation

The theory of the propagation of radio waves has been greatly developed since the last meeting of the Union, through the work of Larmor[1] on ionic refraction, followed by that of Appleton and Barnett[2] in England, and Nichols and Schelleng[3] in America who have independently developed a theory of the action of the earth's magnetic field on the phase velocity of radio waves in an ionized medium such as the Kennelly-Heaviside layer is assumed to be.

From this it follows that the ionized layer acts in the same manner as a quartz crystal in optical phenomena; that is, there is a rotation of the plane of polarization for transmission in the direction of the magnetic field, and double refraction for transmission at right angles to it. If the carriers in the ionized layer are electrons, as seems most probable, it also follows that there should be an absorption band for a wavelength of about 214 m. A. Meissner[4] in Berlin has looked for this absorption band experimentally, but was unable to discover any large increase in absorption in this

* Original Manuscript Received by the Institute, January 5, 1928.

* Presented at the International meeting of the Union in Washington October, 1927.

Note: References will be found at end of paper.

exact neighborhood. It is known, however, that the daylight range of stations of wavelengths between 150 m. and 250 m. is less than the ranges at considerably longer or shorter wavelengths.

Field Strength Measurements

Absolute field-intensity measurements at distances too great for the use of simple measurement instruments, thermoelements, etc., appear to be among the most difficult in physics. In practically all the methods used a comparison is made between the signal to be measured and an artificial signal from a local high-frequency oscillator. This is true even in the long-wave measurement system of the Bureau of Standards; for though in this case the direct comparison is made with artificial audio signals, the calibration for absolute determinations must be carried out by means of a local radio-frequency oscillator.

The great difficulty in all the various methods used lies in the production of a sufficiently weak known local signal in the antenna for comparison with the signal to be measured. The specific possibilities of error, in addition to the difficulties of definite attenuation of the artificial signal by resistance nets, current transformers, or mutual inductances, lies in the possible presence of false couplings either due to inductance or capacity, which can in general be removed only by careful shielding. It is to such errors in calibration, which often are most difficult to detect, that the large differences (sometimes more than five to one) in the results obtained by various experimenters, are almost certainly due.

It would seem that a direct comparison of the different types of apparatus in measuring the same signal offers the best means of decreasing the possibility of error. Such comparisons[5] have already been made between the long-wave apparatus of the Radio Corporation of America, the Bell Telephone Laboratories, and the Bureau of Standards; and it seems desirable to carry on such intercomparisons of all the types of field-strength measurement apparatus at present in use.

Signals at Moderate Distances (below 1500 km.)—The daylight work by Austin[6] (Brant Rock and Arlington experiments) indicated a regular falling off in field strength with the distance over salt water for wavelengths of 1000 to 4000 m. which could be approximately expressed by the Austin-Cohen transmission formula.

The recent daylight experiments of Hollingworth[7] in England and Scotland on Ste. Assise (FT), wavelength 14,000 m., which

were carried on mostly over land have shown, on the other hand, evidences of interference between ground and reflected waves resulting in well-marked maxima and minima of intensity with increasing distance; so that the receiving field at Aberdeen at a distance of 1000 km. was about three times as great as at Manchester at 650 km.

The measurements which have been made in Washington on the transatlantic stations of the Radio Corporation of America in New Brunswick, N. J., (281 km.), Tuckerton, N. J., (251 km.) Rocky Point, L. I., (435 km.) and Marion, Mass., (660 km.), all of which have wavelengths between 11,500 and 16,100 m., have not shown any definite periodic change of intensity with distance at least in the daytime.[27] At night there is plenty of evidence of interference as the signals from the stations at 435 km. and 660 km. have fallen with considerable regularity far below their day values during the past summer, while those from the two stations at 281 km. have risen slightly at night or remained fairly constant. The conditions of experiments are of course somewhat different from those in England, inasmuch as the signals are measured at one point and the transmitting stations at the various distances transmit on somewhat different wavelengths. Nevertheless, since these experiments, some of which have been continued over many years, indicate a regular falling off of intensity with the distance in the daytime, it would seem that they must be held to be in disagreement with the results of Hollingworth. Therefore it would seem very desirable to carry out further measurements in other places, some of which might well be on signals over water as over land of different characteristics.

The English committee[27] has reported the following in regard to the differences in the behavior of transmitting stations at distances between 300 km. and 1000 km. and between 100 km. and 300 km.

Distances 300 to 1000 km.—(1) No abnormal polarization during daylight in summer. (2) A definite sunset cycle lasting 2 to 3 hours. (3) Night values in general very different from day values. (4) No marked effect of magnetic storms.

Distances 100 to 300 km.—(1) Abnormal polarization always present. (2) No sunset cycle. (3) Night and day values nearly the same. (4) Distinct magnetic storm effects.

The Washington observations[27] on long-wave stations have given the following corresponding results in the same ranges: Between 300 km. and 1000 km. (1) Insufficient observations. (2) Fairly regular sunset cycle of reception. (3) Night values very different from day. (4) Distinct solar and magnetic effects, the same as at shorter distances. Between 100 km. and 300 km. (1) Abnormal polarization usually present day and night. (2) No regular sunset reception cycle. (3) Night values generally not very different from day values, at least in summer. (4) Distinct solar and magnetic effects.

The English and American observations are therefore in agreement on (2) and (3) at the longer distances, while at the shorter distances there is general agreement in all four items. The only positive disagreement is in respect to (4) at the longer distances. In regard to the effect of magnetic storms on transmission at distances of from 300 to 1000 km. the French observations at Meudon on the transmitting stations at Bordeaux and Nantes (310 and 360 km. distant, respectively) are in agreement with the British and at variance with the American observations. Our impression is, however, that the conclusions which can be drawn from the American observations are somewhat less positive in regard to all these statements than those expressed in the British report. There also seems to be greater variability in the signals in America during the summer months than in England, while there is at the same time no very great increase here in variability as winter approaches, such as is indicated by the English report although there is a considerable general increase in strength of signal. The maximum variability in summer may amount to 40 per cent in contrast to the 10 per cent variability of the English observations. Below are given the average percentage variations from the monthly means of WII for June, September, and December, 1925, and these may be compared with the average percentage deviations for the year of three long-distance stations received in Washington. It is seen that the American station at only 281 km. has almost the same variability of day signals as the stations at great distances.

New Brunswick, WII, Distance 281 km.

1925	Variability of 10 A. M. Signals
June	17.6 per cent
Sept.	15.6 per cent
Dec.	19.4 per cent

Average Variability for the year—10A.M. signals

Bordeaux (LY)	19.5 per cent
Ste. Assise (FT)	20.0 per cent
Buenos Aires (LPZ)	19.0 per cent

Another effect, which has been observed in America but apparently not in the other countries of the Union, is the tendency for signal intensities to rise both by day and night when the temperature falls.[8] This has been especially noticeable in the case of long-wave stations at distances under 1000 km. during the extreme cold of the so-called cold waves when the intensity often rises to two or three times its average strength.

Field Intensity Measurements at Great Distances.—Long-continued signal measurements at great distances and on the longer wavelengths have been carried out by a number of observers in recent years. Among the more important of these are the observations of Mesny[9] in France, the Marconi Co.[10] in England, Baumler[11] in Berlin, the Indian Post Office[12] in India, and the American Telephone & Telegraph Co.,[13] the Radio Corporation[27] and the Bureau of Standards[14] in America. The Marconi Co.[10] has also sent an experimental expedition around the world, and the French Navy has carried out measurements on a voyage from France to Tahiti.[15]

This mass of observations, while it has revealed many interesting phenomena and has suggested many interesting problems, has brought out the fact that the results obtained by the different methods of measurement are not by any means always in accord, and that even with the same type of apparatus there are dangers of differences in results when it is used under slightly different conditions.

It also appears that the actual variations in signal intensity from day to day, and month to month renders difficult the comparison of observations which are not continued throughout long periods of time.

Ultra Short Wavelengths.—Ultra short-wave intensity observations can hardly as yet be considered in general to be reduced to a quantitative basis. The Bell Telephone Laboratories[16] have, however, developed measurement apparatus for this purpose which in expert hands is capable of fairly exact measurements.

The changes in the signal intensity with changing wavelength and hour of the day are so great, however, that even with purely qualitative observations much information has been gained regarding the behavior of these high frequencies.

The observations of Taylor and Hulburt[17] covering several years, and the work of various members of the French Committee both in France and in the colonies have thrown much light on the subject. The observers are in general in agreement that the short-wave signals after being heard for a moderate distance from the transmitting station, become inaudible over distances depending on the wavelength, and then at greater distances are again heard with great intensity. The only scientific observations contradicting this have been made by Captain Staut[27] in Senegal. This absence of skip distance also seems to be confirmed by the commercial stations in tropical Africa.

The whole matter of the variation in strength, varying skip distance, etc., with changing wavelength, hour and season, may apparently be explained broadly as due to varying penetration in the Kennelly-Heaviside layer, and its varying height.

Transmission Formulas.—Empirical formulas for transmission on the longer waves have been given by Eckersley[10] (based on the theoretical work of Watson) and by Austin and Cohen[6] (based on the Hertzian expression for the field intensity at some distance from an oscillator with the addition of an empirical exponential absorption term). Fuller[18] and Espenschied, Anderson and Bailey[13] have also given formulas in which different values have been given to the constants in the exponential term of the Austin-Cohen formula.

Since the discovery of the great variability of the signal intensity at different times, the general interest in transmission formulas has been much diminished, as it is evident that any formula laying claim to general accuracy would be so complicated that it could hardly be of practical value even if our knowedge of the subject were sufficient to derive it. The most that can be claimed for any of the formulas thus far suggested is a very rough approximation to the actual results averaged over very long periods. Thus far there has been no attempt to produce a formula applicable to the ultra short waves.

Evidences of Reflection or Refraction from an Ionized Atmospheric Layer

Dr. Smith-Rose and Mr. Barfield[19] have made experiments in England on the angle of the wave front of the received wave. The first experiments, made on very long waves, showed that the

wave front was practically vertical, in agreement with the earlier work of Austin.[20] This may be explained as due to the action of the earth as a reflector, due to its high conductivity, which neutralizes the effect of any horizontal components in the electric field of the arriving wave. When these experiments were repeated at broadcasting wavelengths (300–500 m.), a considerable deviation from the vertical was found.

Observations were also made at the same time on the relative strengths of the signals on vertical and closed coil antennas. These measurements on a 386 m. wave at a distance of 124 km. showed that the downcoming wave might make any angle between 13 deg. and 34 deg. with the vertical. 34 deg. would correspond to a height of the reflecting layer of 88 km., which is in agreement with the estimates from other measurements, while the smaller angles might be due to multiple reflections, or perhaps reflections from another layer.

The measurements also indicate that the intensity of the downcoming wave is of the same order as that of the ground wave.

Further research into the nature of the electromagnetic field at a receiving station has been carried out in England by Appleton and Barnett.[21] Their experiments were based on the fact that fading is greater on a loop than on a vertical antenna, since the horizontal component of a wave coming down at an angle would affect a loop, but would not affect a vertical antenna. From the ratio of the fading effects observed on the loop and on the vertical antenna it is possible to calculate the angle of the downcoming wave. Experiments between London and Cambridge gave an angle of 60–70 deg. with the ground. Experiments between Bournemouth and Oxford, in which the wavelength was gradually changed from 385 m. to 395 m., gave seven successive interference maxima and minima. This would indicate a reflecting layer at a height of from 80 to 90 km. In later experiments photographic records of the signal maxima and minima were obtained. In this case observations were continued throughout the night and showed that the height of the reflecting layer increased gradually from soon after sunset until just before dawn. After sunrise it fell rapidly and the intensity of the downcoming wave gradually diminished until it was inappreciable.

In order to determine the attenuation of the ground wave for comparison with the wave reflected from the ionized layer, day

measurements on the variation of signal intensity with distance have been made in England by G. A. Ratcliffe and M. A. F. Barnett[22] on the transmissions from Daventry (1600 m.) and from 2LO (260 m.). These observations were in close agreement with Sommerfeld's theory of over land transmission for distances beyond 10 wavelengths, though deviations from the theory were found at shorter distances. In the calculations the ground resistivity, previously found by Smith-Rose and Barfield[19] was used. These experiments in conjunction with the fading experiments already described, indicated a refraction attenuation coefficient in the reflected wave of 30 per cent.

Work on the polarization of radio waves has also been carried out recently in America by Pickard[23] and by Alexanderson[24] who, using tilting rods and horizontal and vertical receiving antennas, have found for the shorter ranges of wavelength, that the horizontal wave component at a distance may become much stronger than the vertical component. They have also found that this effect is quite independent of the horizontal or vertical position of the transmitting antenna. Mr. Alexanderson also reports periodic changes in the ratio of the horizontal and vertical fields, as the distance is increased. These correspond to the interference fringes of Appleton.

Extensive experiments on ultra short waves at great distances by Taylor and Hulburt,[17] based on the skip distances of signals at different wavelengths both by day and night, and experiments by Breit and Tuve[25] employing 70 m. modulated continuous waves in daylight at a distance of only 8 miles and using an echo effect, have given further information on the height of the ionized layer.

All the observations taken together indicate that the paths of the waves in the upper atmosphere are lower, the longer the wavelength; that they are lower in the day time than at night; and lower in summer than in winter. Taylor and Hulburt's observations indicate heights for ultra short waves on winter nights of more than 500 km; and for summer nights of roughly 300 km., with the corresponding day values considerably lower. Breit and Tuve find for the 70 m. wave in summer, daylight heights of 80–230 km. Appleton's results in the broadcasting range give night heights of about 90 km. in summer. In a recent letter to *Nature*[26] he reports winter night heights of 250–350 km., falling to 90 km. at daylight. This, he thinks, may indicate a second layer.

Apparent Direction Variations in Radio Signals

The British, French and American committees[27] have all carried on extensive observations covering several years on the variations in apparent direction of transmitting stations.

Effect of Wavelength on the Variations.—Leaving out of account the ultra short waves and considering only those from 300 m. upwards, the English committee reports no certain variations in the amount of deviation due to wavelength, although differences are observed in the behavior of certain stations, which may, however, be due to the character of the ground over which the signals pass. The French and American observations are not entirely in agreement with this, as both have found what has seemed to them sufficient proof of a greater tendency to variation as the wavelength is increased. In this connection the French committee reports that the observations at Meudon show for Bordeaux, distance 510 km., at a wavelength of 23,400 m. general weekly maximum deviations of 20 deg. to 25 deg. in some cases even going to 90 deg.; while at 18,900 m. the weekly maxima have ranged between 10 deg. and 15 deg. In the case of Nauen, distance about 900 km., the 13,000 m. wave gives maximum weekly deviations of 7 deg. to 15 deg. and the 18,000 m. wave 0 deg. to 30 deg. There is, I think, general agreement that slight changes in wavelength often produce great changes in bearing variation.

Effect of Distance.—According to the English observations, for distances up to 15 or 20 miles, night variations are slight and not very different from those observed by day. Above 30 miles overland large variations are observed at all wavelengths, which increase up to about 150 miles. Over sea these respective distances are increased by approximately three times. For still greater distances there seems to be a tendency for the variations to decrease, until at 3000 miles or more, the variations amount to only about 3 deg. both day and night. American observations on very long-wave European stations indicate night variations of 6 deg. to 12 deg. with day variations of 2 deg. or 3 deg. The French report mentions a tendency for night variations on the long waves to continue over considerable periods on the same side of the true bearing.

The Cause of the Variations.—It seems to be now generally accepted that night variations are caused by a wave which comes down at an angle from the upper atmosphere with its wave front

so tilted that the lines of magnetic force cut the top and bottom as well as the sides of the direction-finding coil. Proof of a descending wave of this character has already been given under the discussion of the ionized layer, and recent English experiments have shown that night variation is almost entirely eliminated by a direction-finder composed of four vertical Hertzian rods with connections from the breaks at their middle points to the observing house in the center of the system where they are connected to the crossed primaries of a Bellini Tosi goniometer. This is a development of the Adcock system and is equivalent to a rotating loop without top or bottom.

Experiments of the English committee have also shown that the shape of the transmitting antenna plays no important part in the production of night direction deviation.

Solar Activity, Magnetic Storms and Radio Phenomena.

In America the work of Espenschied, Anderson and Bailey[13] has shown a connection between magnetic storms and transatlantic radio transmission, causing frequently a reduction of signal at night and probably a slight increase in the daytime.

More recently Pickard,[28] using his own observations in the broadcasting band of wavelengths as well as the observations of others, and employing moving and periodic averaging methods for the analysis of his material, has produced striking evidence of the dependence of radio phenomena on solar activity. His work indicates a decrease in night signal with increasing sunspot numbers at broadcasting wavelengths and an increase on the ultra short waves.

Austin,[29] using field intensity observations made at the Bureau of Standards since 1915, has shown that the long wave transatlantic daylight signals when averaged by months follow in a general way the sunspot numbers with the changing eleven-year cycle. This conclusion can be stated only as probable from 1915 to 1922, on account of possible errors in the signal measurements; but since the beginning of 1922, the increase of signal intensity with increasing solar activity seems certain.

In England, a relationship between magnetic storms and signals has been noticed on stations at a distance of 100 km. to 300 km., but not at greater distances. In France no relationship has been observed either with solar activity or magnetic storms.

Some of the questions which have been suggested by the experimental material are:

1. Does transmission from east to west differ from that from west to east, as indicated by the results of the Marconi expedition?
2. Is there a limit in wavelength beyond which transmission over land is practically identical with transmission over water?
3. Does the over water transmission in certain parts of the world differ materially from that in other parts?
4. Is there a difference in transmission along and across the earth's magnetic field?
5. What are the causes of the ionization of the reflecting layer?
6. Do the waves above a certain frequency fail to return to the surface of the earth?
7. There also remains the question of the amount of correlation betweeen radio transmission, solar activity, and variations in the earth's magnetic field; and how this may differ at various wavelengths and in various portions of the earth.

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DISCUSSION ON LONG DISTANCE RADIO RECEIVING MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1925* (L. W. Austin)

B. H. J. Kynaston: G. W. Pickard[1] in commenting on K. Sreenivasan's discussion[2] of Dr. Austin's paper states that the observations

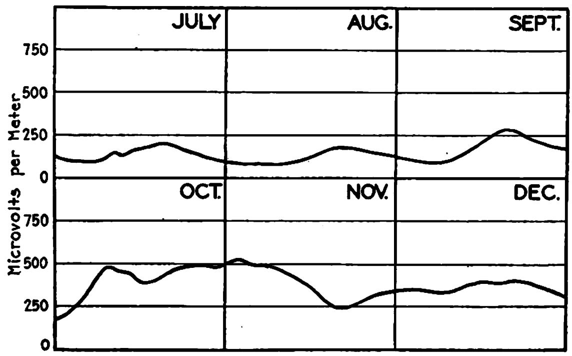

Fig. 1

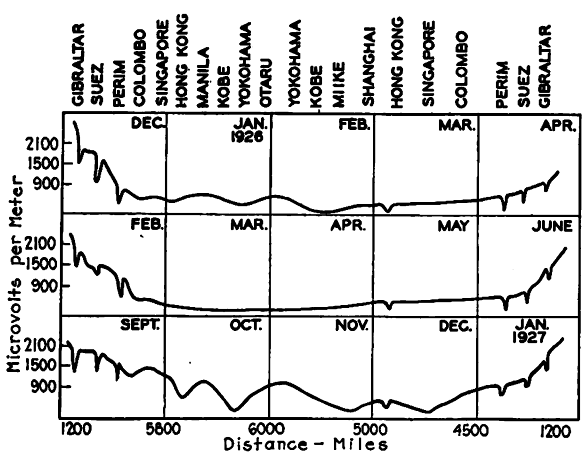

Fig. 2

on Madras (Fort) Radio are too brief for a good comparison with terrestrial and cosmic elements. My own observations taken in the Far East on Rugby's transmissions cover rather longer periods

Original Manuscript Received by the Institute, December 14, 1927.

* Proceedings of the Institute of Radio Engineers, 14, 663: 1926.
[1] Proceedings of the Institute of Radio Engineers, 15, 539: 1927.
[2] Proceedings of the Institute of Radio Engineers, 15, 155: 1927.

than those given by K. Sreenivasan and may perhaps be useful to other investigators.

The curve in Fig. 1 is for moving weekly averages of observations of Rugby's midnight (G.M.T.) transmission on 18,750 metres. It is interesting to note that the curve for October and November shows intensity changes similar to K. Sreenivasan's observations of Madras.

The various curves given in Fig. 2 were taken during three voyages from Liverpool to Japan and back (via Suez) and cover a period of fifteen months, the time and wavelength of the transmission being the same as in Fig. 1. Many of the sudden and rapid changes in the curves were obviously due to screening by nearby land. This was most noticeable as would be expected in places like Gibraltar, Perim, Suez, and in the Formosa Channel between Hong Kong and Shanghai. In all these places large masses of land are on both sides of the ship.

NOTES ON THE DESIGN OF RADIO INSULATORS*

By

T. Walmsley

THE purpose of these notes is to amplify and apply existing knowledge upon the design of some types of radio insulators, used for transmission purposes. There seems to be no general appreciation of the fact that better results can usually be obtained by proportioning insulators correctly than by increasing the quantity of material used. Increased thickness of a dielectric, having a high dielectric constant, frequently causes a reduction in the breakdown voltage of the insulator. In illustration of this contention, the old experiment of introducing a sheet of glass into an air gap between two electrodes is cited. Previous to the intrusion, the air gap successfully resists the application of a certain maximum potential difference, but as soon as the glass plate is inserted, the air space breaks down.

Problems of design may be conveniently considered under the main headings: (1) quality of material, (2) shape and arrangement of material.

Quality of Material

Under the first heading, attention is directed to the mechanical and electrical properties, and the behavior of the material under sustained mechanical and radio-frequency electrical stresses, applied separately and simultaneously.

The quality of porcelain and glass has, during the past few years, been improved considerably. The most recent tests on glass made by the author show that a potential gradient of 8,000 volts r.m.s. per inch at radio frequencies can be taken by a suitable sample for long periods. Similar samples tested by the National Physical Laboratory for cracking compressive loads gave a minimum value of 13.8 tons per sq. in. The samples were in the form of circular disks 3.0 in. diam., 1.62 in. thick, which for the electrical tests had been previously soaked in pure water.

Porcelain shows about the same results as glass, both in compression and under the application of radio-frequency electrical differences of potential. Glass does not appear to stand the test of time so well as porcelain, but as far as the author is aware

* Original Manuscript Received by the Institute, September 21, 1927

exhaustive tests upon different qualities of glass have not been made to decide this point.

The tensile strength of glass and porcelain is somewhat uncertain. Porcelain, in the author's view, is more reliable than glass. Tubes of porcelain having a proof load of 0.6 tons per sq. in. are used at Rugby Radio Station for aerial suspension. The breaking tensile load is known to be considerably in excess of this value.

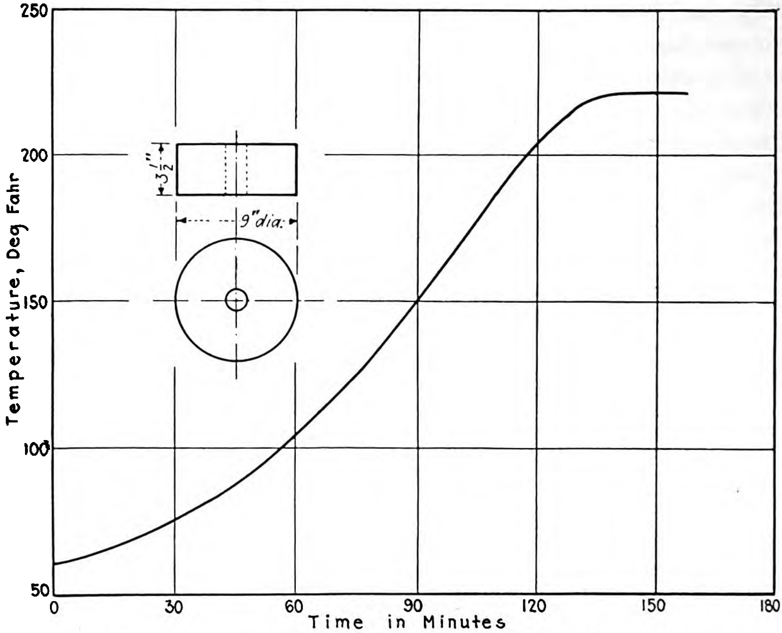

Fig. 1a—Material—Porcelain. Distance between Electrodes—$3\frac{1}{2}$ in. Voltage (R.M.S.)—15,000. Frequency per sec.—50,000. Atmospheric Temperature—50 deg. F.

Makers of a certain propriety type of glass claim a tensile breaking load of over 0.95 tons per sq. in. for their products. The figures, however, are based upon tests onvery small sections. The flexural mechanical strength of suitable types of porcelain is superior to that of glass. In addition to the usual electrical, mechanical, and moisture absorption tests to ascertain the quality of material, a heating test to show the behavior of the material under the application of radio-frequency potential gradients comparable to but more severe than the working conditions of the insulator, is considered essential. It is frequently asserted that a power factor test

made with low voltages under ordinary laboratory conditions, by enabling the energy loss to be ascertained, will yield all the information required. The assumption underlying this contention is that the dielectric constant does not change with increase of potential gradient. It frequently happens, however, when an insulator has not been thoroughly vitrified, that the material is not homogeneous. This is particularly true of solid insulators having minimum dimensions of several inches.

Pockets of insufficiently vitrified material exist within the insulator, surrounded by well-vitrified material. The ordinary power factor test may not disclose any great electrical imperfection, but the application of an intense electric field at radio frequency produces local heating. The moisture of the imperfectly vitrified portion is converted into steam, which, if confined within the boundaries of the well-vitrified walls, causes high mechanical

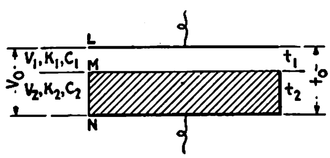

Fig. 1

pressures internally. In consequence the insulator cracks. A typical curve of the temperature rise of an insulator under the application of radio frequency difference of potential is shown in Fig. 1a.

The fall in temperature when the insulator had attained a temperature slightly in excess of that of steam at atmospheric pressure suggests that the vapor in this case was able to escape.

Shape and Arrangement of Material

Several principles are involved in deciding upon the shape of material. Consider Fig. 1. *LM* and *MN* are two elementary condensers in series, having a dielectric constant, K_1 and K_2; capacity C_1 and C_2; thickness t_1 and t_2 respectively.

The potential difference across the two condensers in series $=$ $V_0 = V_1 + V_2$

The potential gradient across

$$LM = \frac{V_0 K_2}{K_1 t_2 + K_2 t_1}$$

When the dielectric is air, the potential gradient across $LM=$

$$\frac{V_0K_2}{t_2+K_2t_1}$$

In the limit when the thickness of air dielectric $=0$, that is, when the tap plate is just touching the solid dielectric, the potential

$$\text{gradient}=\frac{V_0K_2}{t_2}. \qquad (1)$$

This simple expression (1) enables an explanation to be given of the cause of sparking between the surface of a loosely fitting

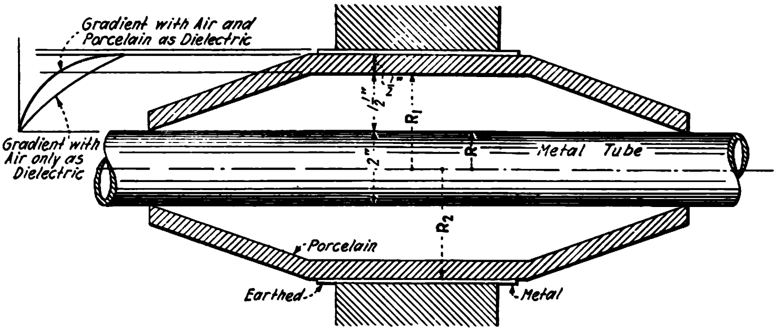

Fig. 2

metal plate and insulator, since the potential gradient across the minute air gap is K_2 times the average potential gradient across the plates. Many cases of breakdown of insulators are attributable to the ionization set up in small air gaps between conducting and insulating surfaces. The cure obviously lies in a correct design of insulator, but the application of graphite to the loosely touching surface is frequently productive of beneficial results.

The type of insulator shown in Fig. 2, frequently found in radio transmitting stations, is bad.

Assuming the material to be porcelain having a dielectric constant K of 5, and regarding the potential difference between internal tube and metal bush as unity, the average potential gradient across the $1\frac{1}{2}$-inch air gap is represented by 0.64 and across the porcelain by 0.078. The average potential gradient in the air gap when the porcelain is removed $=0.56$. Thus the porcelain has caused the average potential gradient in the air gap to be increased

by 15 per cent. The criterion of design is not the average but the maximum potential gradient within the space. This occurs at the surface of the inner conducting tube and is approximately equal to

$$\frac{V_0 K}{R}\left[\frac{1}{K \log_e \frac{R_1}{R} + \log_e \frac{R_2}{R_1}}\right]$$

where R, R_1, R_2 = the radii of inner conducting tube, inside of insulator and outer tube (Fig. 2).

The application of this formula shows that the potential gradient at the surface of the inner conducting tube is increased

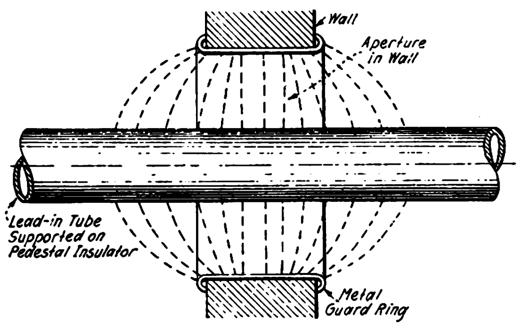

Fig. 3

by about 15 per cent by the introduction of the porcelain bush. Matters could be improved by filling the air space between tube and insulator with a high grade insulating oil, thereby increasing the breakdown voltage and reducing the voltage gradient in the space.

For a lead-in insulator from aerial to transmitter loading coils, an almost ideal arrangement is that shown in Fig. 3.

A similar arrangement for leading high-tension cables into sub-stations and power houses is used on the continent of Europe. Lack of protection against adverse weather conditions is, however, a great disadvantage of this type of lead-in. The aperture might be replaced by a glass or porcelain disk, Fig. 4.

Surface Leakage

This would not greatly disturb the stress distribution. The question of surface leakage, however, must be considered. In the following arguments

R_i = radius of disk on outside diameter of tube.
R_0 = outside radius of disk.
P = a co-efficient denoting surface resistance constant of the material considered.

Then surface resistance of ring δr wide at distance r from center (Fig. 5.) $= \dfrac{\delta r}{2\pi r} \cdot P$

therefore total surface resistance of disk $= \displaystyle\int_{R_i}^{R_0} \frac{\delta r}{2\pi r} \cdot P = \frac{P}{2\pi} \cdot \log_e \frac{R_0}{R_i}$

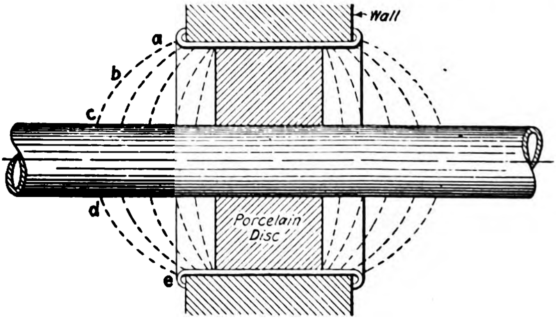

Fig. 4

The correct ratio $\dfrac{R_0}{R_i}$ for maximum spark over voltage in air is about 3. Curve 1 shows the rate at which the leakage resistance increases as the ratio $\dfrac{R_0}{R_i}$ increases from 3 to 27. Thus, although the radius of the disk is increased from 3 to 9 times the radius of the tube, the leakage resistance is only doubled. For a threefold increase in leakage resistance, it would be necessary to have a ratio of $\dfrac{R_0}{R_i} = 27$

In manufacturers' catalogues, leakage length between electrodes is usually quoted. It does not appear to be recognized sufficiently that length alone, without regard to the leakage area, has little value. The subject will be discussed further when the question of

sheds is reviewed. From the figures just quoted two conclusions are formed:

(1) The installation of large-sized glass plates for lead-in windows does not necessarily greatly increase leakage resistance.

(2) An insulator following the contour lines of *abcde* in Fig. 4 will offer a greater leakage resistance than a flat disk of equal diameter.

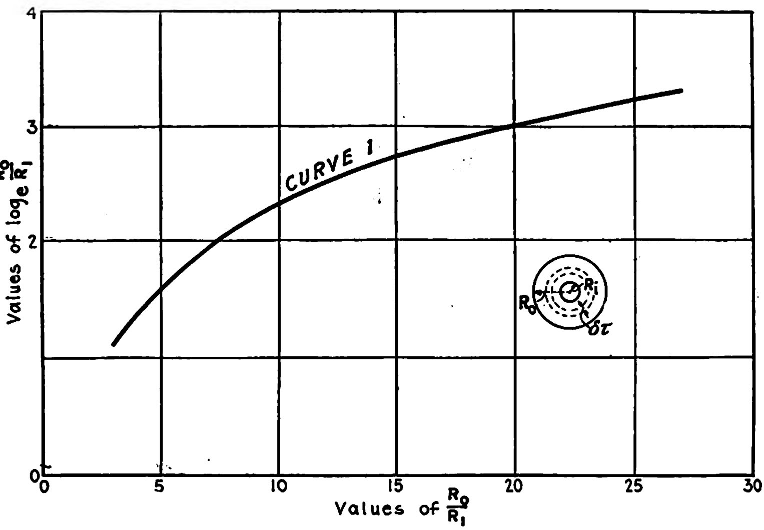

Fig. 5 Curve Showing Relation Between R_o/R_i and $\log_e R_o/R_i$.

Dome Insulators

A natural development from conclusion (2) is the insulator shown in Fig. 6(a), the dome being attached to the walls or glass plate by bolted flanges. A simple insulator which permits a better mechanical arrangement than Fig. 6(a) is that shown in Fig. 6(b). It will be observed that the insulator is placed inside the building. In designing this type of lead-in insulator, the size of tube and aperture would first be decided.

The formulas given by Peek[1] will give sufficiently accurate results:

$$-g_v = g_0\left(1 \div \frac{0.301}{\sqrt{R_i}}\right) \text{ also } g_v = \frac{E}{R_1 \log_e \frac{R_0}{R_i}} = g_0\left(1 \div \frac{0.301}{\sqrt{R_i}}\right)$$

where $g_0 = 30$ kilovolts per centimeter.

[1] "Dielectric Phenomena in High Voltage Engineering."

g_v = gradient in volts per cm. at surface of tube at visible corona.

E = maximum peak voltage between electrodes.

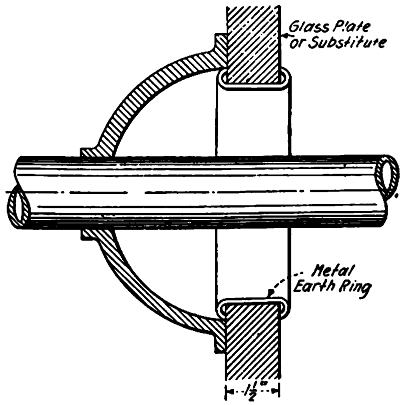

Fig. 6a

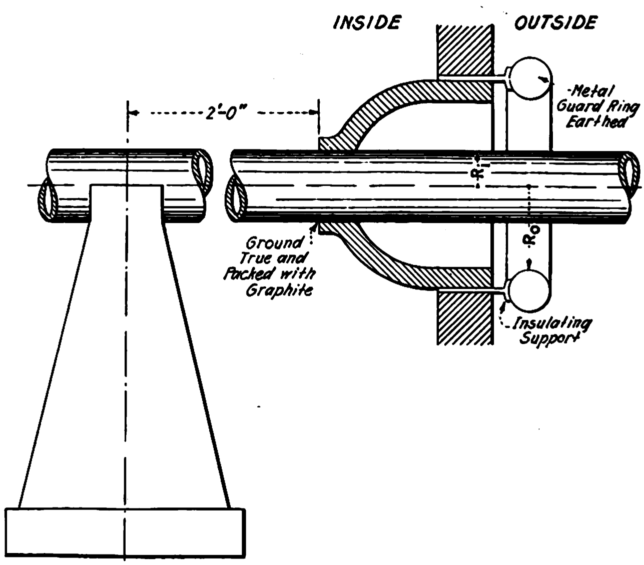

Fig. 6b

SHEDS

Although the ideal insulator has smooth surfaces following the contour lines of stress, sheds are frequently used to increase the leakage length.

The sheds should preferably be part of the main body, and *not* cemented on. Many cases have recently been brought to the

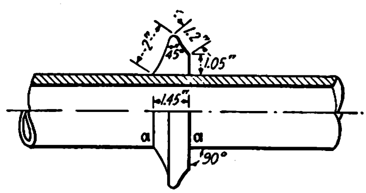

Fig. 7

author's attention of breakages of aerial insulator drip rings due to heating of the cement.

Cement is an imperfect insulator for radio-frequency voltages. In thin layers where the voltage gradient is low it may be used.

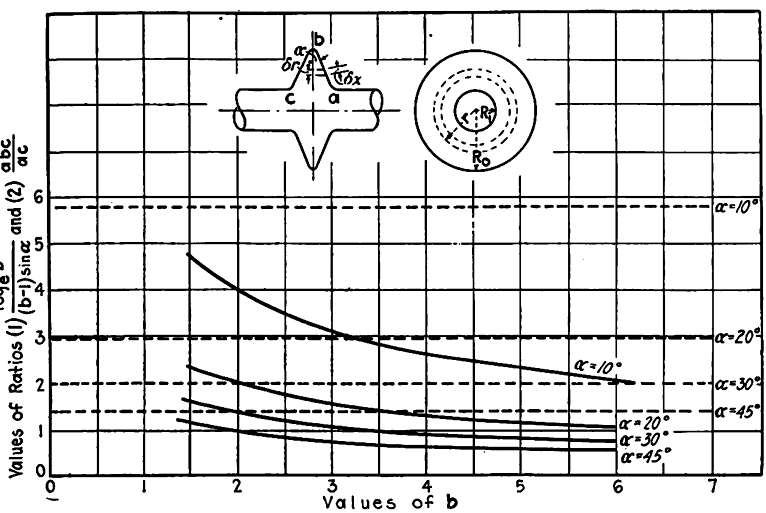

Fig. 8—Curve Showing Relation Between (1) b and $\log_e b\ /(b-1)\sin\alpha$ shown thus ———— (2) b and ratio abc/ac shown thus - - - - - -

For example it is used without ill effect, between the compression insulators supporting the Rugby masts. It connects the metal caps of the aerial supporting insulators to the porcelain

insulator. The metal cap, however, completely encloses the cement, and guard rings protect the insulators. A porcelain drip ring (Fig. 7) cemented along *a a* of a tube insulator will have no such protection and will heat up if subjected to a certain potential gradient. Even though the ring is an integral portion of the body the extra leakage resistance cannot be measured simply by taking the differences between the lengths *abc* and *ac*.

In Fig. 8 a symmetrical ring is shown as an integral part of a tube insulator. To ascertain the value of the surface leakage resistance of the ring in comparison with that due to the plain tube of the same width, i.e., to compare the leakage along *abc* with *ac*, consider a small strip δx

$$\text{Surface leakage resistance of } \delta x = P\frac{\delta x}{2\pi r} = P\frac{\delta r}{2\pi r \cos\alpha}$$

$$\text{therefore total surface resistance of ring } = \frac{2P}{2\pi\cos\alpha}\int_{R_i}^{R_0}\frac{dr}{r}$$

$$= \frac{2P}{2\pi\cos\alpha}\log_e\frac{R_0}{R_i}.$$

Also surface resistance of cylindrical path *ac* with ring removed

$$= P\cdot\frac{ac}{2\pi R_i} = \frac{2P(R_0 - R_i)\tan\alpha}{2\pi R_i}$$

$$\text{therefore } \frac{\text{Surface Resistance of ring leakage path}}{\text{Surface Resistance of cylinder leakage path}}$$

$$= \frac{R_i \log_e \dfrac{R_0}{R_i}}{\cos\alpha (R_0 - R_i)\tan\alpha}$$

$$= \frac{R_1 \log_e \dfrac{R_0}{R_i}}{(R_0 - R_i)\sin\alpha}$$

$$= \frac{\log_e b}{(b-1)\sin\alpha} \text{ where } b = \frac{R_0}{R_i}$$

Curve 2 shows this ratio for various values of b and α. It also shows the *apparent* ratio of the resistance path *abc* to *ac* when no

allowance has been made for the increased area of the ring. It will be observed that in some cases, the presence of sheds actually decreases the value of surface resistance.

For example, when $\alpha = 30$ deg., the surface resistance ratio of ring leakage path and cylinder leakage path when b is greater than

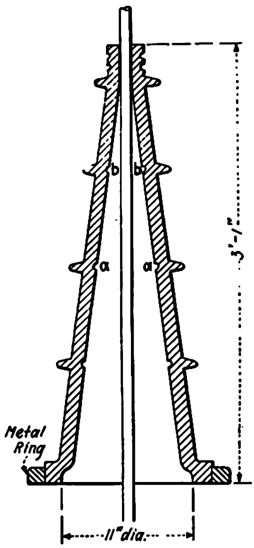

Fig. 9

3.5 is less than unity. It is thus obvious that corrugations on insulators, to be effective, must have a low value of α, i.e., they must be narrow. Further, the ratio $b = \dfrac{R_0}{R_t}$ must be low. It may, of course, be argued that sheds and corrugations keep portions of the insulator dry in wet weather. The author's experience, however, is that fog and snow are more troublesome than rain. Moreover, tube insulators supporting aerials are almost horizontal. In this position the rings offer little shelter from rain, although they break up the stream of water that tends to flow along the tube. This ability to break up the stream is a dubious advantage since it prevents natural washing of the tube. Sheds on dome insulators, the main body of which follow closely the lines of electric stress, should be avoided if possible. Their effect is to produce local

concentrations of electric stress which may cause flashover. In this connection some account of tests made upon a pedestal porcelain insulator by the author, for the purpose of ascertaining the effect of grooves are of interest. A steel tube $\frac{7}{8}$ in. external diameter was passed through the neck and along the axis of the insulator (Fig. 9). The broader extremity of the insulator was surrounded by a metal cap, earth connected. An increasing difference of potential at 16,000 cycles per second was then applied between the bar and cap. When about 50,000 volts r.m.s. had been reached brushing around the groove *aa* inside the insulator was observed. This finally resulted in an arc over at 54,000 volts r.m.s. *inside* the insulator, the arc travelling down the insulator to the metal cap.

The internal $\frac{7}{8}$ in. pipe was then covered by a steel pipe $1\frac{1}{4}$ in. external diameter and a difference of potential again applied. Brushing was observed around the groove *bb*, higher up the insulator than previously, but a flash-over did not occur until 90,000 volts r.m.s. (127,000 volts peak value) had been reached. The breakdown was between rod and guard as far as could be observed. The tests illustrate two main facts: (1) the necessity of correctly proportioning the size of axial conductor to diameter of guard ring, (2) the necessity of making the insulator with ungrooved internal walls. Due to its location within porcelain, the groove was subjected to an increased electric stress.

This would account for the corona. The insulator, previous to the tests, had been soaked in water for several hours.

THE MEASUREMENT OF CHOKE COIL INDUCTANCE*

By

C. A. Wright and F. T. Bowditch

(Research Laboratory, National Carbon Company, Inc.)

Summary—*The investigation described has emphasized the facts that:*

(1) The inductance of the choke coil depends upon the degree to which its core is magnetically saturated because of direct current flowing through its winding.

(2) With a given direct current flowing through the winding of a choke coil, the inductance varies to a marked extent with the magnitude of the alternating current flowing through the winding. Methods of measurement which do not take into account or measure the magnitude of the alternating current are, therefore, unreliable.

(3) The inductance for given conditions may be determined from the saturation curve of the coil. It is determined by the average slope of the saturation curve over the range within which the current varies.

Three modifications of the ammeter-voltmeter method of measuring inductance are presented:

(1) The circuit used in the first modification of this method is applicable where only a few approximate measurements are to be made and where simplicity of connection is of the greatest importance.

(2) The second modification of the method involves simplicity of connection and permits of greater accuracy of measurement than does the first modification, but where a large number of measurements are to be made, it involves inconvenience of manipulation.

(3) The third modification of the ammeter-voltmeter method of measuring the inductance of choke coils involves the use of apparatus not always available, but permits of accuracy of measurement and convenience of manipulation.

Introduction

THERE has been a considerable amount of discussion recently concerning the measurement of the inductance of the choke coils used in *B* power circuits. The methods employed by various investigators do not give comparable results. For example, one company manufacturing choke coils proves by measurements that one of its coils has an inductance of 25 henrys while carrying 50 milliamperes of direct current, while another such company reports a value of about 12 henrys for this same coil, carrying the same direct current. Other investigators report other values, differing from the above and from each other.

The present study of the measurement of choke coil inductance has therefore been undertaken in order to develop a reliable method suitable for the comparative rating of such coils. This paper presents three modifications of such a method, together with a series of measurements exemplifying them. The reasons for the variations noted above are also developed.

* Original Manuscript Received by the Institute, December 7, 1927.
* Presented at Meeting of Cleveland Section, December 2, 1927.

The Function of a Choke Coil in a Filter Circuit

A choke coil consists of a number of insulated turns of wire, wound around and insulated from an iron core. Such a coil may be designed to present a very high impedance to the flow of alternating current, while it provides simultaneously a comparatively low resistance to the flow of a steady direct current. For example, one well-known choke coil under certain typical conditions offers an impedance of over 15000 ohms to the passage of 120-cycle alternating current, and, at the same time, has a direct-current resistance of only 350 ohms.

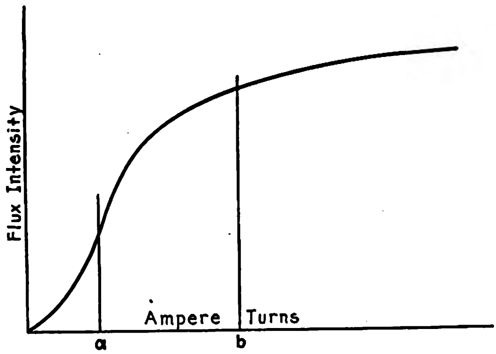

Fig. 1—Typical Magnetization Curve of Transformer Iron.

The choke coil is, therefore, particularly useful in filter circuits in which it is desired to separate the direct current from the alternating current in the pulsating current delivered by a rectifying source. By providing a shunt path of low alternating current and high direct-current impedance (such as the shunt condensers in a filter circuit) the alternating current is induced in large part to take the lower resistance shunt path, while the direct current passes through the choke coils to the load. The higher the inductance of these choke coils, and the higher the capacity of the shunt condensers, the more perfect is this separation of alternating from direct current.

The impedance of a choke coil to alternating current is approximately (the resistance of the coil being neglected) equal to $2\pi fL$, in which f is the alternating current frequency and L is the inductance, which is dependent upon the geometry and character of the iron core and the number of turns and character of the currents flowing in the winding. It is this inductance, L, effective

under the current conditions existing in the filter mesh, which determines the filtering or "choking" value of the coil for the given frequency. The determination of L for a given coil, and the influence of various factors on it therefore constitutes the problem of this investigation.

The Saturating Effect of Direct Current

It is a well-known fact that one of the most important factors affecting the inductance of choke coils is the saturation of the core by the direct current flowing in the winding. Fig. 1 shows a typical saturation curve, in which the magnetizing forces are plotted as abscissas, and the resulting fluxes as ordinates. The magnetizing force for a given coil is directly proportional to the current flowing through the winding, and the resulting flux is dependent upon the number of turns and the nature, size, and shape of the iron core. When the value of magnetizing force (or coil current) exceeds a given amount, the iron core becomes "saturated," and increases in magnetizing force above this point produce but little additional flux. In general, the larger the core and the better its material magnetically, the greater is the direct current which is required to produce this state of saturation.

The inductance of the coil, and therefore, its choking action, is determined by the magnitude of the flux changes produced by the alternating current flowing through it. In other words, *it is determined by the average slope of the saturation curve over the range within which the current varies.* The zero point about which this a-c. variation occurs is determined by the value of direct current which flows through the coil. Thus, if the direct current which fixes the zero point about which the current varies, has a value a in Fig. 1, at the center of the steepest part of the curve, the inductance for moderate values of alternating current will be a maximum. If, however, the direct-current component is of sufficient magnitude to bring the zero point above the knee of the curve (as at b) the inductance is much lower. It is thus apparent that the magnitude of the direct current carried by the coil has a great influence on the effective inductance.

The Effect of the Magnitude of the Alternating Current on the Inductance

A fact which has been overlooked by many investigators and which is responsible for much of the disagreement among them is

that the magnitude of the alternating as well as that of the direct current component affects the inductance of the coil. As the inductance is determined by the average slope of the saturation curve within the limits of a-c. variation, it is apparent that on any but the straight part of the curve, and particularly in the region of the knee of the curve, this average slope is determined largely by the range of the a-c. fluctuation about the zero point. For very low or high values of alternating current, the average slope and consequently the inductance may be lower than it is for intermediate values. It is important in any method of measurement, therefore, that the magnitude of the alternating current as well as the magnitude of the direct current be considered. *In*

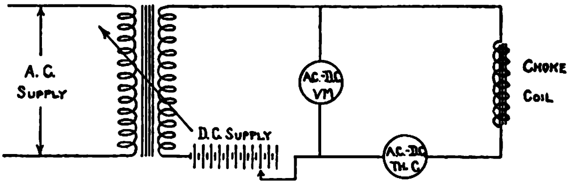

Fig. 2—The Ammeter-Voltmeter Method of Measuring Choke Coil Inductance; First Circuit Modification. (For use where simplicity of connection is important.)

other words, the currents, both alternating and direct, which flow through the coil while it is being measured must be of a magnitude comparable to those which flow through it in the circuit in which it is to be used. Otherwise a false rating is obtained which has no practical application to the problem at hand. Actual measurements taken on a filter circuit of a typical *B* power supply unit show that the first choke coil carries about five milliamperes of alternating current in addition to the direct current furnished to the load. The combination of 5 milliamperes (RMS) of alternating current and 50 milliamperes of direct current through the choke coil has therefore been chosen in this investigation as a standard in measuring coil inductance for comparative purposes.

It is in failing to take account of the magnitude of the alternating current that many bridge, ammeter-voltmeter, three-voltmeter, substitution, and fluxmeter methods of inductance measurement are unreliable. In many cases, the value of this a-c. component is not known, and it is frequently very low. In this investigation, however, there have been developed three modifications of the ammeter-voltmeter method which necessitate the

measurement of the alternating current, and in which it, as well as the direct current, may be adjusted to any values desired. A discussion of these methods follows.

First Modification of the Ammeter-Voltmeter Method of Measurement

The circuit used in the first modification of the ammeter-voltmeter method is shown in Fig. 2. It is applicable where only a few approximate measurements are to be made, and where simplicity of connection is of the greatest importance. Observations are made as follows: With the a-c. source open-circuited at the transformer primary, sufficient B batteries are connected in series with the secondary of the transformer, a thermocouple ammeter, and the coil under test, to produce the desired amplitude of direct current in the coil. The direct voltage required and the direct current are read by means of the a-c., d-c. voltmeter and the thermocouple ammeter. Then sufficient alternating voltage is applied, by means of the variable voltage transformer, to increase the reading of the thermocouple ammeter materially, and the meters are again read.

The inductance calculation is then made in the following manner. When both direct and alternating currents flow simultaneously through an a-c., d-c. ammeter, a deflection is produced which is equal to $\sqrt{I_{ac}^2+I_{dc}^2}$. Similarly, when both alternating and direct voltages are simultaneously applied to an a-c., d-c. voltmeter, the deflection is equal to $\sqrt{E_{ac}^2+E_{dc}^2}$. Having determined, therefore, both the d-c. and the combined deflections, it is possible to solve for E_{ac} and I_{ac}. Knowing that $E_{ac}=(2\pi fL)\times I_{ac}$ the inductance L of the choke coil may be readily obtained by the solution of the equation in which L is now the only unknown quantity. The inductance value so obtained is the value when the measured direct current is flowing through the coil. The resistance of the choke coil is ordinarily neglected in the computation, since it is low in comparison with the reactance and, in addition, is added vectorially to it.

Second Modification of the Voltmeter-Ammeter Method of Measurement

A disadvantage of the method of Fig. 2 is that in order to obtain differences in current readings sufficient to permit of ac-

curacy in the computed vector differences, comparatively large values of alternating current must be passed through the coils, probably larger than those met with in actual practice. This disadvantage may be avoided by connecting in parallel with the thermocouple ammeter a circuit consisting of a variable resistance, a choke coil, batteries, and a d-c. ammeter, as shown in Fig. 3. If the variable resistance and batteries are so connected in parallel with the thermocouple ammeter that they will send a current through it in the reverse direction, and if the batteries and variable resistance are properly adjusted, all of the direct current may be by-passed around the thermocouple ammeter so that it will not interfere with the measurement of the alternating current passing through the circuit. If the impedance of the choke coil in this shunt circuit is sufficiently large in comparison with the impedance

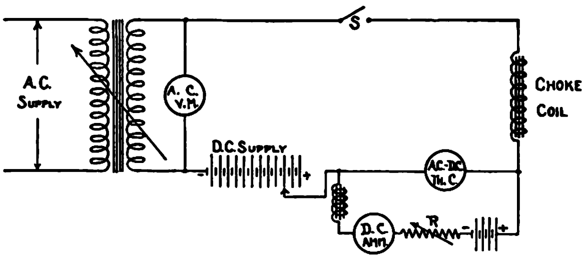

Fig. 3—The Ammeter-Voltmeter Method of Measuring Choke Coil Inductance; Second Circuit Modification. (For use where accuracy rather than simplicity of connection is important.)

of the thermocouple ammeter, the alternating current will not be by-passed around the thermocouple ammeter by this shunt circuit, and will be read accurately by that meter.

If the resistance of the a-c. ammeter is low, an observation is made in the following manner: First, with the a-c. source open circuited, and with the switch S also open, the rheostat R (in the shunt circuit) is adjusted until the ammeters read the desired value of direct current. Then, still leaving the a-c. circuit open the switch S is closed, and the voltage of the d-c. source is adjusted until the reading of the a-c., d-c. ammeter is reduced to zero. The reading of the d-c. meter, however, will be undisturbed, reading the initially chosen direct current which is now being entirely by-passed around the a-c., d-c. instrument.

This latter meter may now be replaced by a more sensitive one, or, if a multi-scale instrument is being used, the switch is

thrown to a more sensitive range. The a-c. voltage is now applied and the variable transformer is adjusted to give the desired alternating current through the choke coil. The current is read on the a-c., d-c. ammeter, undisturbed by the direct current component in the circuit. The a-c. voltmeter is connected directly across the a-c. source as shown, where it reads only the a-c. voltage since the d-c. resistance of the transformer winding is so low that no appreciable direct voltage drop is impressed on the meter. Thus, with both alternating voltage and current readings taken separately from the d-c. readings, the accuracy of the method is good.

If, however, the resistance of the a-c., d-c. ammeter is so high as to form an appreciable part of the total resistance in the circuit

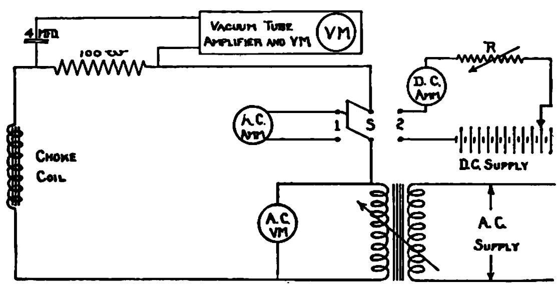

Fig. 4—The Ammeter-Voltmeter Method of Measuring Choke Coil Inductance; Third Circuit Modification. (For use where ease of manipulation and accuracy are important).

(as is likely to be the case if a high resistance thermocouple ammeter is employed) it will be necessary to modify the procedure slightly in order to obtain the desired d-c. balance. A short-circuiting switch is provided around the meter, which is closed initially. The resistance R is adjusted (with the switch S, Fig. 3, open) to give the desired current through the d-c. meter. Then S is closed, and the battery voltage is adjusted until a second d-c. meter, connected in the main circuit, reads exactly the same current. The short circuiting switch around the a-c., d-c. or thermocouple ammeter is now opened, and, if the balance is not perfect, both the rheostat R and the d-c. source must be simultaneously adjusted. Such procedure is necessary only in extreme cases, however, in which it is desired to hold the direct-current constant at a certain definite magnitude, and where a low value of alternating current necessitates the use of a high resistance am-

readily to manipulation, and permits the adjustment over a wide range of both the a-c. and d-c. components.

Inductance Measurements

In the curves plotted as Figs. 5 and 6, certain characteristic data have been chosen for presentation. Fig. 5 illustrates the variation in inductance of a choke coil with the magnitude of alternating current flowing through it. Curve *A* shows such a curve when no direct current is present, indicating that any inductance rating from 30 to 50 henrys might truthfully be claimed

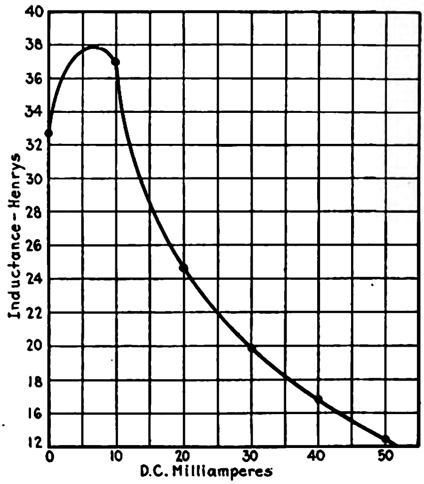

Fig. 6—Variation in Inductance of a Typical "B" Circuit Choke Coil with the Magnitude of Direct Current.
Frequency 60 cycles; 5 milliamperes of alternating current.

for this coil, provided no mention was made of the alternating-current magnitude. This increase in the inductance with current corresponds to the lower bend in the saturation curve shown in Fig. 1. As the magnitude of the a-c. variation about the zero point is increased, the average slope of the included portion of the saturation curve likewise increases with a corresponding rise in the value of the inductance. Curve *B* illustrates the same variation for the same coil, but with 50 milliamperes of direct current flowing. The inductance values are decreased by the saturating effect of the direct current to almost a third of the values in curve *A*, but the increase with the increase in magnitude of the alternating cur-

rent is still apparent. If the magnitude of the alternating current is not recognized in the measurement of this coil, therefore, values ranging from less than 12 to over 20 henrys might be claimed as its effective inductance in a filter circuit carrying 50 milliamperes of direct current.

Fig. 5 shows the variation in inductance of a typical choke coil with the magnitude of the direct current, the alternating current being held constant at 5 milliamperes. Starting from zero, the inductance rises at first, as the zero line of a-c. fluctuation is advanced to the point of maximum slope of the saturation curve (at *a* on the curve of Fig. 1.) As the direct current is further increased beyong this point which gives a maximum value, the knee of the saturation curve is approached, and the inductance decreases as a result. The effectiveness of this coil in a filter circuit would therefore be a maximum if it carried but 6 or 7 milliamperes of direct current, and it would be less than four-tenths as effective if it carried 50 milliamperes of direct current.

It is because of this saturation at high direct currents that a suitable air gap is included in the magnetic path of a well-designed choke coil. Referring to Fig. 1, a similarly plotted curve for air alone would be a straight line, passing through zero, and rising with increasing ampere turns. The slope of this line is less than that for the iron alone at the point *a* but considerably greater than the slope above the saturation point as at *b*. The combined saturation curve for the iron and air gap together is a proportional addition of the two curves, decreasing the slope of the iron over its steepest range, but materially increasing the slope above the saturation point where the iron is usually "worked" in a *B* filter circuit. The effect of the air gap, therefore, is to increase the inductance of the coil over the range within which it is to be used, even though it does decrease the maximum value obtainable at lower direct-current loads.

Inductance measurements showing the variation in inductance of various types of choke coils with the alternating-current frequency were made. In making these measurements, a substitution method was employed in the following manner. In parallel with the choke coil and ammeter, a variable non-inductive resistance and a second similar ammeter were connected. The output of an audio-frequency oscillator was then connected across the terminals of the two parallel circuits, and the variable resistance in

meter. Having obtained the d-c. balance, the remaining operations are made exactly as before.

It is apparent that the method of Fig. 3 has the advantages of accuracy and simplicity of connection, but the disadvantage of difficulty of manipulation.

Third Modification of Ammeter-Voltmeter Method of Inductance Measurement

The method of inductance measurement of Fig. 4 is a special adaptation of the ammeter-voltmeter method which is accurate and which provides convenience of manipulation. It is to be preferred when a sensitive amplifier and vacuum-tube voltmeter are available. Observations are made in the following manner: With the switch S in the No. 1 position, the alternating voltage is adjusted to give the desired alternating current through the choke coil as registered on the a-c. ammeter connected across the switch terminals. The d-c. source, connected across the other end of the switch, is not in the circuit and only alternating current flows through the meter and choke coil. A portion of this alternating current flows through the input transformer of the vacuum-tube amplifier, and produces a deflection proportional to the alternating current flowing through the choke coil. From the value of alternating voltage and current so obtained, the inductance of the coil with no direct current flowing through it may be calculated.

The switch S is now thrown over to the No. 2 position, removing the delicate a-c. ammeter from the circuit and superimposing the direct current upon the alternating current already flowing in the circuit. The rheostat R is now adjusted to give the desired value of direct current, indicated on a D'Arsonval type d-c. ammeter which is not affected by the a-c. component of current which it carries. The alternating voltage is next adjusted to give the same reading on the vacuum-tube amplifier as before, and this new value of alternating voltage, together with the value of alternating current initially chosen, is used to compute the inductance of the coil with the measured direct current flowing through it. In making this calculation, the series resistances in the circuit are ordinarily neglected, since they are low in comparison with the coil reactance, and, in addition, are added vectorially to it.

The reading of the vacuum-tube amplifier is not influenced by the magnitude of the direct current flowing in the circuit and

furnishes a convenient method of adjusting the alternating current to the accurately measured initial value flowing when there was no direct current in the circuit. The method lends itself

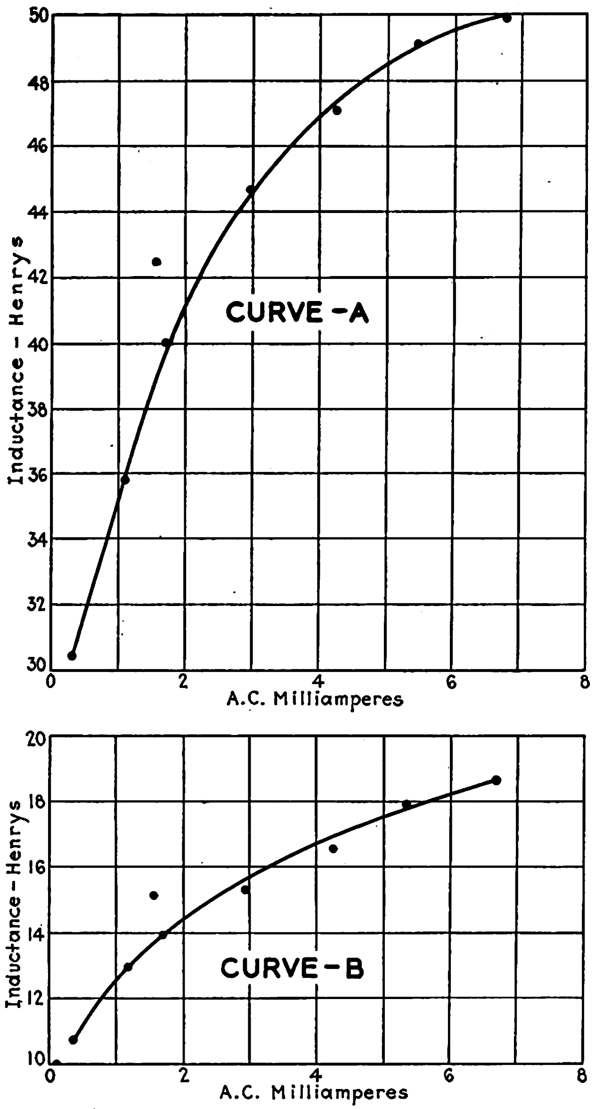

Fig. 5—Variation in Inductance of a Typical "B" Circuit Choke Coil with the Magnitude of Alternating Current.

Frequency 60 cycles. Curve *A*—with no direct current. Curve *B*—with 50 milliamperes of direct current flowing.

one was adjusted until the currents in both branches were identical. The magnitude of the resistance in ohms is then equal to the impedance, $2\pi fL$, of the choke coil in the parallel circuit, and the inductance L may be readily calculated.

The inductance values obtained at 60 and 120 cycles were in close agreement in all cases. Since 120 cycles is the predominant frequency in double-wave rectification, it is thus permissible to make measurements using the much more convenient 60-cycle source with the assurance that the values so obtained will be a reliable indication of the operation of the coil in a filter circuit. There seems to be a slight tendency for the inductance to rise at the higher frequencies, probably due to the increase in eddy current and hysteresis losses.

Although most of the measurements have been made of the inductance of choke coils to be used in radio B filter circuits, the methods have also been satisfactorily applied to choke coils designed for use in radio A filter circuits.

BOOK REVIEW

Applied Magnetism, by T. F. Wall. D. Van Nostrand Company, 268 pages, cloth bound. Price $8.00

This book should serve admirably either as a textbook for engineering students or as a reference book. The first part, "The Principles of Applied Magnetism," includes chapters on "Magnetic Theory," "Definitions," "The Characteristics of Magnetic Substances." Common methods for magnetic testing of materials. The author is the head of the Electrical Engineering Department of the University of Sheffield, England.

It is an easily read presentation of magnetic principles although very little data has been included that cannot be found elsewhere. Both parts take up permanent magnet steels and transformer sheets. Other chapters are devoted to the electron theory of magnetism, magneto-striction, and the generation of intense magnetic fields (the order of half a million gauss). The author's method of generating such fields should prove of interest to the workers of many lines of research.

R. R. Batcher

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The membership of the Institute is asked to refer to page 236 of this issue and supply information relative to the publication of Patent Digests. Answers to the following questions will be appreciated by the Board of Direction:

1. Shall the Institute discontinue the publication of any Patent Digests?

2. Shall the Institute continue to publish Patent Digests such as have been published during the past year?

3. Shall the Patent Digests be published to include one illustration, or drawing, plus one or two claims?

4. Shall the Institute publish a short summary, or abstract of each radio patent?

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 16 April, 1928 Number 4

CONTENTS

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Chairman
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Secretary-Treasurer
A. Crossley, Naval Research Laboratory, Bellevue, D. C.

Laurens E. Whittemore

VICE-PRESIDENT OF THE INSTITUTE, 1928.

Laurens E. Whittemore was born in Topeka, Kansas, August 20, 1892. He was educated at Washburn College, Topeka, Kansas (A.B. degree in 1914) and University of Kansas from which he received the M.A. degree in 1915.

Mr. Whittemore was an instructor in the Department of Physics of the University of Kansas from 1915 to 1917. He left the University to join the Radio Laboratory Staff of the U. S. Bureau of Standards where he remained until 1923. From 1923 to 1924 he was Secretary of the Inter-Departmental Radio Advisory Commission of the U. S. Department of Commerce.

Since 1925 Mr. Whittemore has been an engineer in the Department of Development and Research of the American Telephone and Telegraph Company in New York City.

Mr. Whittemore was appointed to the Board of Direction of the Institute in 1926 and has served as Chairman of the Committee on Standardization of the Institute from 1926 to date. He is a Fellow in the Institute.

He is the author of several Bureau of Standards publications and was the joint author of the "Lefax Radio Handbook."

CONTRIBUTORS TO THIS ISSUE

Ballantine, Stuart: Born at Germantown, Pennsylvania, September 22, 1897. Radio amateur 1908– ; operator, Marconi Co. summers 1914–15, H. K. Mulford Co., bacteriologists, 1916; Bell Telephone Co. of Pennsylvania, 1917; Expert Radio Aide, U. S. Navy, in charge of research and development of radio direction-finder apparatus, Philadelphia Navy Yard, 1917-20; organized Philadelphia Section of the Institute, 1920 and served as Chairman until 1926; studied mathematics Drexel Institute, Phila., Pa. 1919, and mathematical-physics in Graduate School, Harvard University, 1920-21; with L. M. Hull organized research work at Radio Frequency Laboratories, Inc., Boonton, N. J., 1922–23; John Tyndall Scholar in Mathematical Physics, Harvard University, 1923–24; privately engaged in miscellaneous research work in radio, spectroscopy, astrophysics, propagation of electric waves in the upper atmosphere, at White Haven, Pennsylvania, 1924–27; in charge Research Division, Radio Frequency Laboratories, Inc., 1927– . Mr. Ballantine has been a frequent contributor to the PROCEEDINGS, and is an Associate member of the Institute.

Cady, W. G.: Born at Providence, Rhode Island, December 10, 1874. Received Ph.B. degree, 1895, M.A. 1896, Brown University; Ph.D., 1900, University of Berlin. Instructor in mathematics, Brown University, 1895-1897. Magnetic observer, U. S. Coast and Geodetic Survey, 1900–02. Instructor in physics, 1902–1903, associate professor, 1903–1907, professor physics since 1907, Wesleyan University, Middletown, Connecticut. Dr. Cady has been elected a member of the Board of Direction of the Institute of Radio Engineers for the year 1928, is a member of the Meetings and Papers Committee, and has frequently contributed technical papers to the PROCEEDINGS. He is a Fellow of the Institute.

Crossley, Alfred: Born at Newark, New Jersey, June 2, 1892. Special course in radio engineering at University of North Dakota under Dr. A. Hoyt Taylor. Associated with various radio activities since 1910 in U. S. Navy, Tropical Radio Company, and Dupont Company. During world war was Lieutenant (j.g.) U. S. Navy, in research work on special trans-oceanic receiving sets and submarine radio equipment. In 1919 became Expert Radio Aide U. S. Navy having duties in connection with direction of naval radio research. Since 1923 associated with Naval Research Laboratory doing work in connection with frequency standardization, piezo-electric crystal control and receiving sets. He is a Member of the Institute of Radio Engineers and of the American Society of Naval Engineers.

Dreher, Carl: Graduated from the College of the City of New York, 1917, having specialized in electric and radio engineering courses. From 1917 to date has been associated with the Marconi Wireless Telegraph Company of America, The General Electric Company, the Radio Corporation of America, and National Broadcasting Company. From 1923 to 1927 was Engineer-in-charge of Stations WJZ and WJY of the Radio Corporation of America. Is now Staff Engineer of the National Broadcasting Company, conductor of the "As the Broadcaster Sees It" Department in *Radio Broadcast*; director of the Radio Club of America and business manager of its *Proceedings*; expert examiner in radio for the Municipal Civil Service Commission of New York. He is a Member of the Institute of Radio Engineers.

Hanna, C. R.: Born at Indianapolis, Indiana, December 17, 1899. Received the B.S. degree in E.E. from Purdue Universitv in 1922 and the E.E. degree in 1926. Since July of 1922 he has been associated with the Westinghouse Electric and Manufacturing Company. Since January, 1923 he has been in the Research Department of this organization working on radio and acoustic problems.

Hoch, E. T.: Received the B.S. degree in E.E. from the Case School of Applied Science in 1914; with the Western Electric Company, Manufacturing and Installation Department, 1914–1915; Engineering Department of the Western Electric Company, 1915–1924; Bell Telephone Laboratories, Inc., 1925 to date. Engaged in dielectric studies and condenser development, and in high-frequency coil design.

Sutherlin, Lee: Born in Putnam County, Indiana, July 11, 1889. Graduated from Central Normal College, Danville, Indiana, with B.S. degree, 1909; studied at Indiana University, Bloomington, Indiana, 1909–1912; received the A.B. degree in 1912; University of Chicago during the summers of 1915–1916–1917, received the M.S. degree in 1917. Head of Physics Department Muncie High School, Muncie, Indiana, 1912–1916. Instructor in Applied Mathematics, Culver Military Academy, Culver, Indiana, 1916–1917. Served as enlisted man and officer in Signal Corps, U. S. Army, December 12, 1917 to September 9, 1919. Radio engineer with Miller-Reese-Hutchison, Inc., New York City, 1919–1920. Radio research engineer, Westinghouse Electric and Manufacturing Company, October 1920 to date. He is an Associate member of the Institute.

Terman, Frederick Emmons: Born at English, Indiana, January 7, 1900. Received the A.B. degree from Stanford University, 1920; E.E. degree, Stanford University, 1922; Sc. D. Massachusetts Institute of Technology, 1924. Now Assistant Professor of electrical engineering at Stanford University, and in charge of communication and analytical work. He is an Associate member of the Institute.

Terrell, William D.: Born at Golansville, Virginia, August 10, 1871. Entered telegraph service in 1889; became radio inspector, Department of Commerce, July 1, 1911 at New York City; transferred to Washington, D.C. and placed in charge of the Radio Division, March 1, 1915, in which capacity he has served to date. Charter associate member of the Institute.

Upp, Charles B.: Born in Highland County, Ohio, March 5, 1893. Graduated from grade schools, Rainsboro, Ohio, and high school, Greenfield, Ohio. Received the B.S. in E.E. degree from Ohio State University, 1919, and the M.S. degree in 1923. During the world war served in the Coast Artillery Corps. From 1919 to 1922 was an engineer with the General Electric Company Fort Wayne, Ind. in transformer design. From 1923 to date has been in the Research Department of the Westinghouse Electric and Manufacturing Company.

Walsh, Lincoln: Born November 3, 1903. Foreman, manufacturing, J. Walsh & Brother, Elizabeth, New Jersey, 1919–1921. With Bell Telephone Laboratories in summers of 1923–24–25. Received the M.E. degree Stevens Institute of Technology, 1926. Engaged in receiver development work, Hazeltine Corporation, 1926–1927. 1927 to date, consulting engineer. He is an Associate member of the Institute.

Wheeler, Harold A.: (See Proceedings for January, 1928.)

INSTITUTE ACTIVITIES

March Meeting of the Board of Direction.

At the meeting of the Board of Direction of the Institute held on March 7, 1928, in the offices of the Institute, 33 West 39th Street, New York City, the following were present: Alfred N. Goldsmith, President; Melville Eastham, Treasurer; Ralph Bown, Junior Past President; Donald McNicol, Junior Past President; Arthur Batcheller, W. G. Cady, J. H. Dellinger, R. A. Heising, R. A. Manson, R. H. Marriott, and J. M. Clayton, Secretary.

The following were transferred or elected to the higher grades of membership in the Institute: Transferred to the grade of Fellow: A. A. Oswald, J. C. Schelleng, W. Wilson. Transferred to the grade of Member: F. A. Cobb, A. C. Hofmann, E. M. MaDan, F. J. Marco. Elected to the grade of Member: Quinton Adams, F. H. Amis, C. G. Cadman, J. P. Johnston, W. L. Krahl, Cecil McQuillan, L. C. Herndon.

One hundred and six Associate members and sixteen Junior members were elected.

Licensing of Radio Engineers

The following letter from Everett N. Curtis, Counsellor-at-Law and a member of the Institute, regarding the requirement of the laws of the State of New York for licensing of practicing radio engineers is printed for the information of the membership of the Institute:

February 18, 1928.

The Institute of Radio Engineers,
33 West 39th Street,
New York City.

Gentlemen:

Replying to yours of the 15th inst., I beg to advise you that, in my opinion, a member of the Institute, who is practicing or offering to practice radio engineering in the State of New York, either in a consulting capacity or as an engineer in a manufacturing organization, is required, under the law relating to Engineers and Surveyors, to submit to the State Board of Licensing evidence that he is qualified so to practice and to obtain a certificate of license to practice before he can lawfully engage in professional engineering. Under Section 1466 a person practices "professional engineering" where he holds himself out as able to do, or who does, the work that an engineer does in the planning, designing, constructing, inspecting and supervising of engineering work or appliances involved in public or private projects, or in making investigations for proposed engineering projects.

Under Section 1463 a non-resident of the State of New York is exempted from the operation of the Law, where he has no established place of business

in the State, and where his practice does not exceed thirty days in any calendar year, provided he is legally qualified in his own state or country. He is also exempted for a reasonable time, where he has filed an application for a license and paid the fee. There are also exempt, under Section 1463, employees or pupils acting under the direction of a licensed professional engineer, and not in responsible charge or supervising as principal; officers and employees of the United States acting solely as such; and the practice of professional engineering solely as an officer or employee of a corporation engaged in interstate commerce.

Very truly yours,
(signed) Everett M. Curtis

Research Fellowships at Wisconsin

Two research fellowships in Engineering are to be appointed on April 30th by the University of Wisconsin. Candidates must be graduates of engineering colleges of recognized standing, and, preferably, should have had one or two years of graduate study, of teaching, or of engineering experience. Applications will be received up to April 15th. Information and application blanks can be obtained from Dean F. E. Turneaure, College of Engineering, Madison, Wisconsin.

The appointments will be for a period of two years, subject to satisfactory service, and the salary will be $900 for the first year and $1100 for the second year. A fellow will be expected to devote not less than half time to assigned research in the College of Engineering, but will be given an opportunity to complete the requirements for a master's degree within the two-year period. The period of service will be the usual academic year, including the short vacations.

Changes of Address

On page XXVI of this issue of the Proceedings, in the advertising section in the rear, will be found a list of members, mail for whom has been returned to the offices of the Institute on account of incomplete addresses or changes of locations. It will be helpful if members of the Institute knowing the present address of any persons listed therein will notify this office of their current address so that copies of the Proceedings may reach them promptly.

Institute Meetings

New York Meeting

At the New York meeting of the Institute, held on March 7, 1928 in the Engineering Societies Building, 33 West 39th Street,

a paper by C. R. Hanna, L. Sutherlin, and C. B. Upp of the Westinghouse Research Laboratory, entitled "Development of a New Power Amplifier Tube" was presented by Mr. Sutherlin.

The paper is published in this issue of the PROCEEDINGS.

In the discussion which followed the presentation of the paper the following, among others, took part; Messrs. Sutherlin, Crom, Batsel, Hull, Goldsmith, and Herbst. Over three hundred members and guests attended this meeting.

BOSTON SECTION

A meeting of the Boston Section was held in Cruft Laboratory on March 16th. Dr. G. W. Pierce presented a paper on "Control of Radio Audio Frequencies by Means of Magneto Striction."

On March 30th there will be a joint meeting between the local section of the American Institute of Electrical Engineers and the Boston Section of the Institute to hear a paper by Dr. E. J. Burg, of Union College, Schenectady, N. Y.

BUFFALO-NIAGARA SECTION

On February 15th in Foster Hall, University of Buffalo, a meeting of the Buffalo-Niagara Section was held. Dr. L. G. Hector presided.

Carl Dreher, staff engineer of the National Broadcasting Company, presented a paper "Problems of Broadcast Operation." The paper is being published in this issue of the PROCEEDINGS.

Messrs. Eichman, Hector, Pickett, and others participated in the discussion which followed.

Forty-three members of the Institute attended the meeting.

CHICAGO SECTION

On January 30th, in the Monadock Building Auditorium, Chicago, C. W. Horn, of the Westinghouse Electric and Manufacturing Company, presented a paper, "Some Short Wave Experiences." J. H. Miller presided.

This was a joint meeting with the local section of the American Institute of Electrical Engineers, the Western Society of Engineers, and the Chicago Section of the Institute.

The attendance at this meeting was over one hundred and thirty.

On February 17th a meeting of the Chicago Section was held in the auditorium of the Western Society of Engineers. J. H. Miller presided.

Professor R. R. Ramsey of Indiana University presented a paper "Radiation from Aerials."

Messrs. Miller, Wilcox, Kranz and other participated in the discussion.

Twenty-four members of the Institute attended.

Cleveland Section

A meeting of the Cleveland Section was held on March 2nd in the Physics Building of the Case School of Applied Science. Professor John R. Martin presided.

Two papers were presented. The first, by George H. Mills, instructor in electrical engineering, Case School of Applied Science, was on "The Electrodynamic Loudspeaker and Associated Power Amplifier."

The various sources of distortion in radio reproduction were discussed, such as the pick-up and broadcasting at the station, in the receiver, and especially in the loudspeaker. The shortcomings of the horn type of speaker were pointed out and comparison made to the cone type. It was stated that the cone type of usual commercial design would handle more volume without distortion than the usual horn type.

The type of driving unit used in the two types were discussed. The usual horn type uses an iron armature drive in many cases, similar to the usual cone type—both having their limitations as to the volume they can handle. The later moving coil or electrodynamic driving unit has many advantages over the other type of drive. A pronounced advantage is its ability to handle great volume without distortion.

The details of construction—mounting of moving coil, suspension of cone, and design of field coil—were pointed out. Diagrams of Kolster and R.C.A. 104 Amplifiers, together with moving coil speakers, were shown and explained. An interesting point was the use of the speaker field as a choke coil. A Kolster Speaker and Power Amplifier was demonstrated both on phonograph records with a magnetic pick-up and on broadcast reception with a receiving set.

The second paper by C. B. Hamman, of the Department of Physics, East Technical High School, was a review of a recent paper "A New High Efficiency Horn Type Loudspeaker," published in rhe Bell *Technical Journal.* This talk was confined entirely to a teview of the paper in which the speaker used lantern slides made

from illustrations in the *Journal.* The speaker described the use of the electrodynamic type of driving unit in combination with an exponential horn. The moving coil is attached to a small corrugated metal diaphragm. The energy is transmitted to the air column in a very novel manner.

The above papers were discussed by Professor Martin and Messrs. Worden, Leonard, and Kintner.

On April 6th there will be a meeting of the Cleveland Section in the Case School of Applied Science. Dr. Miller and Professor Martin of the Case School will present a paper on "Loudspeaker Analysis with the Phonodike."

Connecticut Valley Section

On February 20th a meeting of the Connecticut Valley Section was held in the auditorium of the Hartford Electric Light Company. Dr. K. S. Van Dyke presided.

Professor C. M. Jansky, Jr. presented a paper on "Some Studies of Radio Broadcast Coverage in the Middle West."

On February 29th a meeting of the Section was held in the auditorium of the Hartford Electric Light Company. Dr. W. G. Cady presided.

Professor Hidetsugu Yagi of Japan presented a paper "Beams of Ultra Short Waves." The paper explained the directional properties of waves over the use of reflectors or directors, or both, and defined these terms. Many curves were displayed showing the received energy from different locations of the sending and receiving antenna with and without reflectors and directors for both vertical and horizontal polarized waves.

Polar diagrams were also displayed showing the energy distribution with reflectors both with and without directors. Conclusions drawn were that any radiating system now devised can be improved upon by the use of wave directors.

The second part of the paper dealt with the use of a magnetrone for the generation of radio frequencies from twelve to forty centimeters wavelength. Circuit diagrams were shown, and curves giving the effect of anode voltage and magnetic field were explained.

It is hoped that this paper can be published in full in a forthcoming issue of the Proceedings.

Fifty-two members of the Section attended this meeting.

Detroit Section

Dr. Phillip Thomas, of the Research Department of the Westinghouse Electric & Manufacturing Company, presented a paper before the Detroit Section of the Institute on February 24th in the West Engineering Building of the University of Michigan, Ann Arbor.

The paper, entitled "Radio and the Transmission of Power by Radio," showed by demonstrations the close analogy between mechanical and electric vibration and standing waves, using a pendulum, a vibrating rope, and an oscillating system giving about thirty-five watts at a wavelength of 2.4 meters. After showing how small and simple such apparatus becomes on waves of 2 meters or less, Dr. Thomas discussed the difficulties incident to generation of very short wavelengths, and described a line of attack which it is hoped will result in the generation of power of the order of kilowatts associated with a twenty-centimeter wavelength. The possibility of ionization of the air by a beam radiated at this wavelength, and some useful applications of such a "weightless wire" were touched upon. Dr. Thomas concluded his address with the statement that his company hoped shortly to produce such a wave and test its characteristics.

In the discussion which followed it was brought out that the energy density in this contemplated beam of radiation would be about twenty times as great as that of sunlight, at high noon, at the earth's surface. E. T. Glatzel presided at the meeting.

Over four hundred and twenty-five members and guests attended the meeting. Thirty-five persons attended the dinner preceding the meeting.

Los Angeles Section

An informal dinner preceded the meeting of the Los Angeles Section held on February 20th in the Elite Cafe, 633 S. Flower Street, Los Angeles. Thirty-three members attended the dinner.

The new feature of Section meetings, "Timely Topics," consisted of a short paper on the Radio Ray read by A. P. Hill.

J. Clement, of San Diego, gave a short technical description of KFSD.

E. W. Butler of Cunningham, Inc. presented a paper on vacuum tubes covering the tubes in use today.

The latter paper was freely discussed by all members present. Sixty-five members attended the meeting.

Philadelphia Section

A meeting of the Philadelphia Section was held on February 24th in the Bartol Laboratories. J. C. Van Horn presided.

The paper of the evening by Knox McIlwain and W. S. Thompson was entitled "Radio Field Strength Survey of Philadelphia."

The paper was discussed by Messrs. Babcock, Wilson, Van Horn, and others.

Following the technical session, the election of 1928 took place. J. C. Van Horn was re-elected Chairman of the Section and J. C. Mevius, Secretary.

San Francisco Section

The San Francisco Section held a meeting on February 15th in the rooms of the Engineers' Club, 206 Sansome Street, San Francisco. Dr. L. F. Fuller presided.

George T. Royden, of the Mackay Radio Company, presented a paper on "The Development of the Kolster A-C. Radio Receiver." The paper was accompanied by a demonstration of the operation of the receiver and included a technical description of the constructional and engineering features.

A general discussion on the part of the fifty-five members present followed.

Seattle Section

On February 25th in the Club Room of the Telephone Building, Seattle, a meeting of the Seattle Section was held. W. A. Kleist presided.

A motion picture film showing the apparatus employed in television was presented. This film gave an indication of the equipment required and an idea of the method used in accomplishing television.

L. D. Robinson presented a paper on "Transformer Design" in which the fundamental or primary characteristics of transformers were discussed and the method of cutting the laminations and annealing the iron was illustrated. The exciting current was discussed and mention was made of the harmonics which resulted. Core losses were analyzed and methods of reducing them were discussed.

A second paper by T. M. Libby and W. A. Kleist on "The Importance of Various Frequency Bands in the Voice and Music Spectrum" demonstrated by means of a set of phonograph records the importance of these bands. Records were prepared by elimi-

nating certain bands of frequencies by means of electrical filters. It was pointed out that certain of the records, from which all frequencies except a narrow band in the middle of the range had been eliminated, sounded very much like the radio sets of a few years ago.

Washington Section

A meeting of the Washington Section was held on March 8th in Picardi's Cafe, 1417 New York Avenue, Washington, D.C. Dr. A. H. Taylor presided.

A paper by L. C. Young and Dr. A. H. Taylor on "A Study of Short Wave Propagation with special reference to Round the World Signals" was presented by Dr. Taylor. It is expected that this paper will be published in a forthcoming issue of the Proceedings.

Messrs. Dellinger, Whittemore, Robinson, Pratt, Blair, Stewart, Merryman and others participated in the discussion which followed.

Fifty-three members and guests attended the informal dinner which preceded the meeting and sixty attended the meeting. Following the technical session the election of 1928 officers took place with the following results:

Chairman—F. P. Guthrie

Vice-Chairman—Dr. C. B. Jolliffe

Secretary-Treasurer—Alfred Crossley

The next meeting of the Washington Section will be held on April 12th at Picardi's Cafe.

Institute Committees

Committee on Meetings and Papers

A meeting of the Committee on Meetings and Papers was held in the offices of the Institute on March 6th at 2 p.m. The following members were present: J. H. Dellinger, Chairman; W. R. G. Baker, M. C. Batsel, Zeh Bouck, W. G. Cady, E. T. Dickey, Carl Dreher, Edgar Felix, W. G. H. Finch, H. A. Fredericks, V. M. Graham, C. R. Hanna, Sylvan Harris, D. G. Little, E. L. Nelson, G. W. Pickard, W. C. White, W. Wilson, and Messrs. Marriott and Clayton.

The general plans and policies covering the activities of this Committee for the ensuing year were discussed. It is planned that the work of the Committee for 1928 will, in the main, be handled

through correspondence. Various plans looking to increased participation in discussions of papers presented before meetings were outlined.

Several suggestions were brought forth tending toward more intimate contact with the Meetings and Papers Committees of Sections.

Committee on Membership

A meeting of the Committee on Membership was held in the offices of the Institute on the evening of March 1st. The following were present: H. F. Dart, Chairman; F. R. Brick, J. M. Clayton.

The "Aims and Activities" booklet of the Institute was revised for 1928 and a number of matters relative to increase in membership were discussed.

Committee on Sections

On March 13th at 2 P.M. a meeting of the Committee on Sections was held in the offices of the Institute. The following members of the Committee were present: Donald McNicol, Chairman; E. R. Shute, Quinton Adams, M. Berger, Arthur Batcheller, Harvey Klumb, and H. C. Gawler.

Material for a booklet to be used in the guidance of members interested in the organization and operation of Sections, as submitted by the 1927 Committee on Sections, was approved and is to be transmitted to the Board of Direction with the recommendation for adoption and printing.

Various other important matters regarding the operation of Sections were discussed.

Committee on Admissions

At the meeting of the Committee on Admissions of the Institute held on March 7th, the following were present: R. A. Heising, Chairman; E. R. Shute, Louis M. Hull, Frederick Vreeland.

The Committee considered twenty-five applications for admission or transfer to the higher grades of membership, acting favorably upon fourteen of them. A number of applications had to be held over to the next meeting of the Committee due to the quantity on hand for consideration.

Personal Mention

C. H. Nordhaus is now in the Engineering Department of Grigsby-Grunaw-Hinds Company of Chicago.

J. M. Davidson, formerly with the African Theatres, is now head of his own radio engineering organization at Salisbury, Rhodesia.

George T. Royden, Radio Engineering Department of the Mackay Radio and Telegraph Company, has been transferred from San Francisco to Honolulu.

B. A. Halfpap has resigned from the Radio Corporation of America and is now connected with the Federal Railway Institute of Milwaukee, Wisconsin, in the capacity of Radio Instructor.

V. Ford Greaves has resigned from the staff of the Magnavox Company of Oakland, California, to become associated with Newcombe-Hawley, Inc. of St. Charles, Ill., in the Engineering Sales Division.

W. S. Fithian has recently become associated with the Victor Talking Machine Company of Camden, New Jersey. Mr. Fithian was formerly in the Supply Department of the Philadelphia Electric Company.

THE INTERNATIONAL RADIOTELEGRAPH CONFERENCE OF WASHINGTON, 1927*

By

W. D. Terrell

(Chief of Radio Division, U. S. Department of Commerce)

The International Radiotelegraph Conference which held its sessions in Washington from October 4th to November 25th of this year was composed of delegates from 79 countries including those colonies and possessions whose delegates affixed separate signatures to the treaty which was drawn up. There were nearly 300 government delegates and in addition about 75 representatives of communication companies and other interested international agencies.

The specific purpose of the conference was to revise the International Radiotelegraph Convention or Treaty which was signed at London in July, 1912. It may be stated broadly that the general purpose of the conference was to formulate such provisions as are appropriate for international agreement, in order to minimize interference between radio stations engaged in international service or which are international in their interfering capabilities.

The delegation of the United States to this conference consisted of 15 men commissioned for this duty by President Coolidge. These delegates came largely from government departments, but there were also several from private life and from the communication companies. Herbert Hoover, Secretary of Commerce, was Chairman of the American Delegation and President of the Conference. It is interesting to note in passing that the chairmen of the two largest and quite important delegations were engineers, Herbert Hoover, Chairman of the American delegation, and Col. T. F. Purves, Chairman of the British Delegation. Col. Purves is the Chief Engineer of the British General Postoffice.

The conference was organized into committees under the chairmanship of delegates from the various countries. The conference was provided with the necessary staff for the publication and distribution of proposals, the preparation of reports and

* Original Manuscript Received by the Institute, December 21, 1927.
* Presented at the Annual Institute Convention, January 9, 1928.

A detailed account of the activities of the Conference published in book form can be obtained from the United States Government Printing Office at Washington, D. C. for 40 cents, the exact title being "International Radiotelegraph Conference, and General and Supplementary Regulations Relating Thereto."

making of arrangements for entertainment, and the conduct of other business of the conference. Officers of the International Bureau of the Telegraph Union coöperated with the American Delegation on matters of this kind.

It had been customary at previous International Radiotelegraph Conferences to conduct all proceedings in the French language alone. Through the efforts of the American Delegation, an agreement was reached whereby English was used for the discussions as well as French. Interpreters were provided by the American Delegation. The official texts of the convention, regulations and other documents were in French. English texts were prepared and distributed by the American Delegation.

The treaty which was drawn up consists of three parts known as the Convention, the General Regulations, and the Supplementary Regulations. The Convention is based largely on that which was signed at London in 1912. The Regulations are based to some extent on a draft of regulations for mobile radio service which was formulated at an Interallied Conference in Washington in 1920, the technical features having been revised at an Interallied Conference in Paris in 1921.

The Department of State of the United States Government had submitted to the International Bureau at Berne, Switzerland, in 1926, the American proposals for the convention and regulations and had forwarded in addition the proposals of American Radio Companies for management regulations or operating rules which would be the subject of agreement between the radio operating agencies engaged in international communication. The decision to divide the regulations into two parts, General Regulations and Supplementary Regulations, was made by the conference at the request of the United States. Those provisions which are of a managerial nature and relate to the operation of radio services, particularly those operated by private companies, could in this way be put into the supplementary regulations which would not be signed by the delegates from the United States. This supplementary set of regulations also contains a provision making applicable to radio the provisions of the International Telegraph Regulations to which the United States is not a party. No difficulty is involved in this, however, since the United States did not sign the supplementary regulations. The delegates of Canada and Honduras also refrained from signing the supplementary section.

The provisions of the Convention cover such matters as the licensing of radio transmitting stations and operators, the secrecy of messages, intercommunication between coastal and ship stations, the mutual exchange of information necessary to facilitate international radio communication, the settlement of accounts in the mobile service, the establishment of an international consulting committee on radio communication, the allocation of blocks of call letters to countries, and the settlement of disputes by arbitration. The Convention contains no article on the question of voting. The decisions on the questions before this conference were so nearly unanimous that on only one occasion was a roll call required at a plenary session. It was agreed that the next conference would determine its own rules governing voting.

Perhaps the most important article of the regulations is Article 5 which deals with the allocation of frequency and wavelength bands to radio services. In the allocation waves are designated in the first instance by their frequency in kilocycles per second. Following this designation there is given the approximate wavelength in meters. The conversion factor used is 300,000.

The frequency allocation adopted is not an allocation to countries. It is entirely an allocation to services, the stations of all countries having equal rights to the use of the bands designated for a particular service.

There are several provisions of the regulations which give a tacit recognition to the rights of priority of radio stations which have been in operation—for example, in paragraph 16 of Article 5, it is provided that the frequencies assigned to all new fixed, land or broadcasting stations must be chosen in such a manner as to prevent as far as possible interference with international services carried on by existing stations, the frequencies of which have already been notified to the International Bureau. Another provision of this article requires that notice be sent in advance to the International Bureau of the establishment of stations using frequencies below 37.5 kilocycles per second (wavelengths above 8000 meters) in case the use of this frequency might cause international interference over broad areas. Similar notice must be given the International Bureau regarding the operation of short wave stations intended to carry on a regular service and which are likely to cause international interference.

The frequency allocations to the various services conform in their major divisions to the assignments which have been used in

the United States under the recommendations of the 4th National Radio Conference. The band from 10 to 100 kilocycles (30,000 to 3000 meters) is assigned to stations engaged in point-to-point service, chiefly of course the trans-oceanic service. The band from 100 to 550 kilocycles (3000 to 545 meters) has been designated primarily for ship to ship, ship to shore, and aircraft services. This includes radio beacons on a band at about 300 kilocycles (1000 meters) and provides for a radio compass service on a band around 375 kilocycles (800 meters). The 500-kilocycle frequency (600 meters) is the international calling and distress wave and may be used for message traffic only on condition that interference with call signals and distress signals will not result. The band between 194 and 285 kilocycles (1550 and 1050 meters) was one on which it was somewhat difficult to secure agreement. The difficulty arose from the fact that many of the European countries desired to utilize this band for broadcasting. It was finally agreed that part of this band may be used for broadcasting in Europe only, and that the rest of the band will be used by mobile and aircraft services and by fixed stations not open to public correspondence.

The band from 550 to 1500 kilocycles is now universally recognized as the broadcasting band. Permission is given to use one frequency in this band, namely 1365 kilocycles (220 meters) for small ships. The entire band may be used by mobile service in any part of the world where its use will not interfere with broadcasting.

The band from 1500 to 60,000 kilocycles (200 to 5 meters) has been divided into 40 smaller bands and apportioned between mobile services, communication between fixed stations, broadcasting and amateur stations. This allocation of the short waves involves some change from the 4th National Radio Conference allocation, but has the advantage of giving some assurance that stations of a given type which begin operation in this band will be able to continue in operation subject only to the adjustment of interference with other stations engaged in a similar service.

The Conference gave definite recognition to the amateur in international radio communication. The allocation to amateur service of four exclusive bands and two non-exclusive bands was secured through the efforts of the American Delegation with the support of the delegates from Canada and New Zealand. The result is to give the amateurs much greater assurance of making contact with one another internationally.

While the Conference recognized that the allocation of frequency bands to specific services was necessary in order to minimize interference, there was a corresponding desire to leave to each country or to any group of countries in a certain region as much freedom as possible in making assignments to stations which are not international in their effect. Only international stations therefore must comply with the allocation. Freedom is left for the assignment of any frequency to stations which do not cause international interference.

It was recognized as inadvisable to write into the regulations definite provisions of a technical or engineering nature which might become obsolete during the next few years. Instead, general provisions calling for the maintenance of a high technical standard were adopted. It is provided, for example, in Article 4 of the General Regulations, that the wave emitted by a station must be maintained upon its authorized frequency as closely as the state of the art permits, and its radiation must be as free as practicable from all emissions not essential to the type of communication carried on. The interested administrations shall fix the tolerance allowed between the mean frequency of emissions and the recorded frequency; they shall endeavor to take advantage of technical improvements progressively to reduce this tolerance. The width of a frequency band occupied by the emission of a station must be reasonably consistent with good current engineering practice for the type of communication involved.

In considering the use of damped wave transmitters, the Conference felt that definite dates must be set on which certain restrictions would go into effect. The Regulations provide that upon adoption, no more damped wave sets are to be installed at fixed or land stations; that after January 1, 1930 sets of this type to be installed on ships shall, when working on full power, use less than 300 watts measured at the input of the supply transformer. The use of existing damped wave transmitters is to be discontinued by all land stations on January 1, 1935. It is provided, however, that no restriction shall be placed upon the means which an operator of a mobile station in distress may use of attracting attention, indicate his position and obtain assistance.

The regulations annexed to the London Convention were applicable exclusively to ship-to-ship and ship-to-shore services. In the new Washington Regulations most of the provisions are applicable to mobile service which now includes aircraft as well

as ships. In addition to provisions regarding the use of traffic frequencies other than the calling frequency there are regulations providing for the necessary control of traffic by land stations, the routing of messages by mobile stations and other related matters.

Provision is of course made in the regulations for absolute priority in the mobile service for distress calls, messages and traffic. In addition to SOS as a distress call, there is now established a radio telephone distress call which consists of the spoken expression "may day." This corresponds to the pronunciation in French of the expression which means "help me." Provision is made for the use of an alarm signal which has as its purpose the setting into operation of an automatic apparatus to give an automatic alarm and to warn someone on the ship in which it is installed that a distress signal is about to follow. The alarm signal consists of a series of 12 dashes sent in one minute, the duration of each dash being 4 seconds and the duration of the interval between dashes one second.

Two new urgency signals, *XXX*, and in the case of aircraft *PAN*, are established for indicating that urgent messages are on hand concerning the safety of the ship or aircraft or of persons on board or in sight from them. A safety signal, *TTT*, is established to be used as a preamble to a message concerning the safety of navigation or containing meteorological warnings.

Article 6 of the General Regulations contains provisions relating to the issuance of operator's certificates. These provisions differ but little from the present practice of the United States except for the provision that in order to secure a place as a Chief Operator on a vessel of the first class it will be necessary for an operator to have had a year's experience under a first class license.

Provisions are included in the regulations designating the hours of service of ships having one or two operators respectively. Complete revised lists of abbreviations or *Q* signals are included, applicable to aircraft communication as well as other situations.

Regulations are included applicable to special services such as meteorological services, time signals, notices to navigators, radio compass, and radio beacon service. Rules are included covering the settlement of accounts in the mobile service.

The Conference, throughout its work, endeavored to keep before it the principle that its conclusions should be of such a nature as not to interfere with the development of the art. It is generally

felt that the Convention and Regulations adopted occupy a safe middle ground between avoidance of restriction and the maintenance of orderly communication. The Convention and Regulations become effective on January 1, 1929 for all of the governments which ratify it. The Conference accepted an invitation from the Government of Spain to meet next in Madrid in 1932. Egypt and Holland had also extended invitations.

The outstanding impression left by the Conference is perhaps the fact that every effort was made by all of the delegates to secure the correct solution of the problems under discussion. The technical questions in particular were usually discussed and the conclusions arrived at from a technical rather than a nationalistic standpoint. The general attitude was one of coöperation and of realization that the problems should be solved on their technical merits.

MODES OF VIBRATION IN PIEZO-ELECTRIC CRYSTALS*

By

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Summary—*The presence of nodes and antinodes on the surface of oscillating quartz crystals have been discovered. The symmetrical arrangement of these nodal points permits a study of the modes of vibration in the crystal plate and the use of the following formulas for determination of the velocity of sound waves through quartz and Young's modulus.*

$$V = F2T \qquad e = V^2D.$$

where V is the velocity, F the frequency, T the thickness of the plate, e Young's modulus and D the density. The value obtained for V was 5733 meters per second while 8.785×10^{11} C.G.S. units represents Young's modulus for plane parallel to X-axis dimension.

THE piezo-electric crystal and its application for frequency control of vacuum-tube transmitters has held the attention of radio engineers for the last four years and the major activities along these lines have been devoted to the practical application of this means of frequency control. The practical application of the piezo-electric crystal has been such that no appreciable time has been devoted to the study of modes of vibration in the crystal. In view of this condition there is presented in this paper information which has been obtained at the Naval Research Laboratory which will shed some light on this interesting and important subject. It is hoped that this information will stimulate an interest among investigators in this phase of the piezo-electric crystal art and lead to more knowledge of this phenomenon.

The major part of the data presented in this paper was obtained with zero angle or the "Curie" cut type of quartz crystal when this crystal was oscillating at a frequency which corresponds to the thickness or X axis.[1] The crystals employed in this experiment were approximately 25 millimeters square and of different thickness from 0.6 to 6 mm.

It has been known that it is possible to increase or decrease the piezo-electric response characteristic of a quartz crystal, when oscillating at radio frequencies, by adjusting the several dimensions of the crystal. An increase in piezo-electric response

* Original Manuscript Received by the Institute, Feb. 20, 1928.

[1] "Piezo-Electric Crystal-Controlled Transmitters," A. Crossley, Proc. I.R.E., Jan. 1927.

is obtained when the several dimensions are integrally related to each other, while a decrease is obtained when this condition is departed from.

The proper dimensioning of crystals for maximum piezo-electric response was thought to have a definite relation to the phenomenon of resonance from a mechanical and acoustical standpoint. This was found to be true and evidence was obtained which showed the presence of nodes and antinodes on the surfaces of the crystal. The presence of the nodes and antinodes was noted in an experiment which was intended to determine whether or not the vibrating crystal would evaporate water. For this

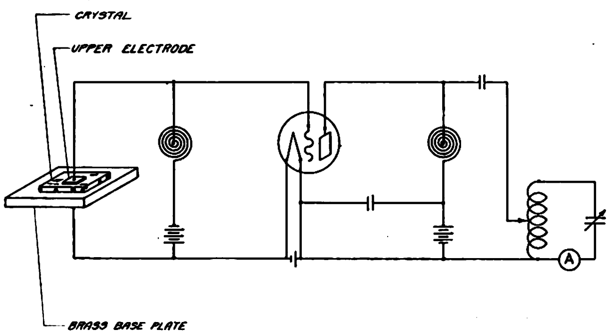

Fig. 1

experiment a crystal-controlled vacuum-tube oscillator of approximately 30 watts output was employed. The crystal was placed on the brass base plate, shown in Fig. 1, and a small brass electrode made contact with the upper surface of the crystal. A portion of the upper surface of the crystal (4000 kcs. fundamental frequency), was exposed and when the system was oscillating a small drop of water was placed on the exposed part of the crystal. As soon as the water came in contact with the crystal the greater part of it rose in a column of vapor while the remaining portion split up into small particles and assumed definite positions, and then disappeared, presumably by a process of evaporation. The various particles referred to arranged themselves in a group of squares, each particle separated from the others by a definite distance.

The water particles were hard to observe, and resort was made to the use of various solutions and suspensions such as ink, potassium permanganate, ferro ferricyanide, cupric ferricyanide, and red lead and alcohol for obtaining a permanent pattern on the

surface of the crystal. Of the different indicators cited the ferro ferricyanide produced the best record. The potassium permanganate when mixed with water evaporated quickly and returned to the crystalline form. Blue ink was fair, but had such a surface tension that it would not readily break up into the particles and separate without leaving clouded areas around the particles.

Following the classical experiments of Chladni and Kundst resort was made to the use of lycopodium powder, flowers of sulphur and red lead to note whether these materials would form patterns on the crystal surface. The lycopodium powder when

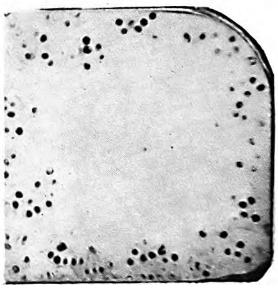

Fig. 2

placed on the oscillating crystal was thrown off and no trace of the powder remained on the crystal. The flowers of sulphur and red lead were treated the same way, but some of the heavier particles remained on the surface of the crystal and became heated. Some particles which were of the right area, say 1 mm. in diameter, started whirling around, and kept whirling until a wind current carried them off the crystal surface.

A small globule of amalgamated mercury was placed on the crystal, and it immediately heated up and started boiling and after a period of time the mercury evaporated and a solid metallic deposit was left on the surface of the crystal. This deposit could not be brushed off and only by grinding could it be removed. It is thought that this deposit was therefore an alloy of copper and zinc, because the mercury had been previously placed in a brass thermometer well.

There are submitted (Figs. 2 and 3) enlarged photographs of the patterns produced on crystals using ferro ferricyanide. When a pattern was obtained on one part of the surface of the crystal the upper electrode or plate was shifted over to another position

on the crystal thus exposing a new surface for another portion of the pattern. This procedure was kept up until the greater part of the surface was covered with the pattern. During the experiment care should be taken so that no liquid will creep in between the crystal and upper electrode, which condition causes the crystal to stop oscillating.

It was not possible to obtain a perfect reproduction of the position of the dots for as soon as the circuit stops oscillating the particles would roll out of line or they would recombine with other particles as can be noted from the photograph. A slight change in the fundamental frequency due to the loading of the

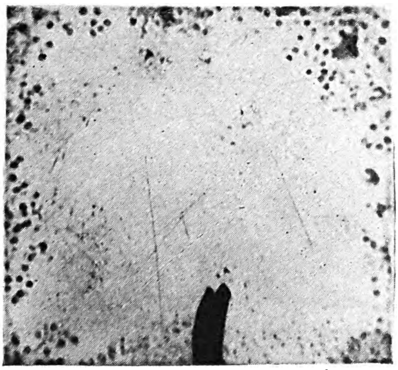

Fig. 3

crystal, which is apparent during the evaporation process, also causes the dots to move away from and back to the original position. This latter condition can be observed very definitely by listening to the beat note obtained from a local heterodyne for as soon as the beat note changes there is a simultaneous change in the position of the dots. This change in the position of the dots in most cases causes a blur or irregular alignment of the dots.

The best reproductions can be observed along the edge of the crystal. A very vivid demonstration of the nodes and antinodes can be had by submerging the crystal in transformer oil to a depth of approximately ¼ inch, and driving the crystal by use of a self-oscillating circuit. When a frequency equivalent to one of the crystal frequencies is impressed on the crystal and part of the surface of the crystal is exposed to the oil a multitude of eruptions are observed. The picture presented on the surface of the oil

resembles a flat plane with numerous little hills situated at definite distances from each other. The hill tops represent point of maximum movement in the quartz plate while the low areas are minimum movement. The oil phenomena is opposite in observed effect to the ink-spot phenomena for in the case of the ink spot only the point of no movement is shown while in the other case the maximum movement is indicated.

A study of these photographs and numerous other ink-spot patterns obtained with different crystals shows us that there are spots of no movement represented by the deposits and places of maximum movement which are indicated by the absence of any kind of a deposit. The distance between nodes is equal to one-

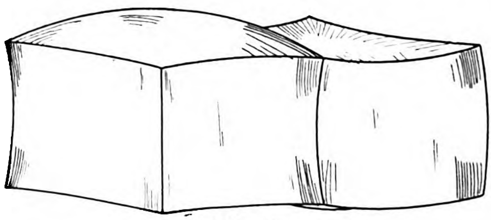

Fig. 4

half a wavelength of sound in the crystal and is also equal the thickness of the crystal. Measurement on two crystal specimens of the average distance between indicated nodal points or ink spot showed that this distance checked to within less than one per cent of the thickness of the crystal. This check is of great interest as it ties in very accurately with our integral relation law.

A further study of results obtained during these experiments suggests the following explanation of the mechanism of oscillation or vibration observed in the crystal. Let us assume from a vibration standpoint that the crystal is of a mosaic form, having numerous crystal cubes which when assembled make up the crystal plate. Each cube will have all three dimensions equal to the thickness of the crystal plate, which we also can assume is equal to one-half a wavelength of sound in the crystal. Now, let us examine the sketches shown in Figs. 4 and 5, which are intended to show graphically the distortion which takes place in the respective suggested cubes of the crystal plate when the plate is oscillating. It is, of course, understood that the distortion is greatly exaggerated in these sketches for the purpose of explaining the vibration phenomenon. In Fig. 4 two cubes have been drawn from data

obtained by the summation of the vector forces due to stresses in the three dimensions and also from data obtained from the crystal patterns previously described. The figure upon inspection suggests a flexural vibration. Sufficient data are not on hand to state definitely whether or not it is a flexural or longitudinal vibration.

We note from these sketches when the surface of one cube is rising the surface of the next cube is falling like that of sections of a tent which are subjected to gusts of wind. We also note that the boundary points such as the four corners of the upper and lower surfaces of the cube are points of no movement, which checks with our observed deposits at the nodal points of the crystal. Fig. 5 shows the same two cubes as depicted in Fig. 4, but

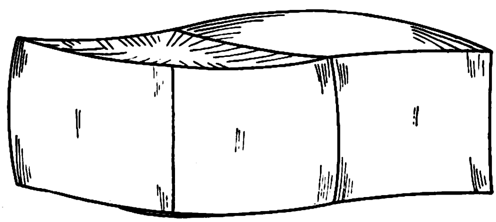

Fig. 5

under the influence of the opposite mechanical force, therefore, these two figures show the two extreme mechanical distortions which exist in a crystal when oscillating at one of its natural periods.

Another way to consider this oscillating phenomenon in the crystal is to consider the crystal as a checker board which has the alternate red and black squares on its surface. In such a form let us visualize the instantaneous oscillation effect by considering for one moment that the black squares are moving upward while the red squares are moving downward, and the next moment the red squares start moving upward while the black squares are moving downward. The maximum movement is at the center of each square tapering off to no movement at the corners as shown in the sketch.

The whirling of the particles of sulphur and red lead previously mentioned can be easily explained by the fact that there is a spot of no motion at the corner of the squares and the alternate rising and falling of the portions of quartz surrounding this point can impart the whirling or circular movement to the particle.

Assuming that the suggested phenomena of the oscillating condition of the crystal is correct it is possible to obtain data on the velocity of sound waves through quartz and also Young's modulus or the coefficient of elasticity of quartz. The velocity of sound waves through quartz was obtained by measuring the frequency and the thickness for the X-dimension of 18 crystals ranging in thickness from 0.6 to 6 mm. and averaging the values obtained by use of the following formula:—

$$V = F\lambda \text{ or } V = F2T$$

where V is the velocity, F the frequency, λ the wavelength of sound in quartz and T is the thickness for the X-dimension. The thickness of the crystal as previously stated is equal to one-half wavelength of sound in quartz for the X-axis oscillation. The average value obtained for the velocity is 5733 meters per second.

Knowing the velocity and assuming the density to be equal to 2.67 grams per cc. and substituting these values in the following formula we have:—

$$e = V^2 D$$

where e is Young's modulus and D the density. The value of Young's modulus thus obtained is equal to 8.785×10^{11} C. G. S. units.

The value obtained for Young's modulus of quartz in the oscillating state applies only to the zero-angle crystal and for the oscillation corresponding to the X or thickness dimension. Different velocities and Young's moduli will be obtained with 30 degree crystals and also for the longitudinal mode of oscillation. The value of Young's modulus herein quoted lies between values obtained by other experimenters[2] who employed the conventional mechanical method for determining this elastic constant. The latter values are as follows: 10.3×10^{11} C.G.S. units for a plane parallel to the optical axis and 7.85×10^{11} when at right angles to optical axis. Considering the two entirely different methods of measurement of Young's modulus and the small difference noted between the radio-frequency measurement and either of the mechanical values it is reasonable to assume the suggested theory of crystal oscillation merits consideration.

The author wishes to express his appreciation to Drs. L. P. Wheeler, H. B. Maris, Messrs. H. D. Eisenhauer and J. W. Wright for their assistance in this investigation.

[2] Handbook of Chemistry and Physics, page 443.

The following publications are quoted as having some bearing on the subject matter of this paper:

A. Hund. Proc. I. R. E. August, 1926.
A. Crossley, Proc. I. R. E., January, 1927.
A. Meissner, Proc. I. R. E., April, 1927.
E. P. Tawil, *Comptes Rendus*, December, 1926.
E. P. Tawil, *Comptes Rendus*, July 11, 1927.

SOME CHARACTERISTICS AND APPLICATIONS OF FOUR-ELECTRODE TUBES*

By

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Summary—*Four-electrode tubes may be classified by their designs and uses as "screen-grid" tubes, "space-charge-grid" tubes, and "double function" tubes. In the screen-grid tube the inner grid is the control electrode and the outer or screen-grid is kept at a fixed potential. The capacity between plate and control-grid is thereby reduced to an almost negligible value. A second result of the screen-grid is a large increase in amplification factor and plate resistance without reduction of mutual conductance. This permits high amplification in connection with high impedance coupling circuits, without undesired regeneration or oscillation. The screen-grid principle has been applied to transmitting tubes as well as receiving tubes. Characteristics of these tubes are given in detail.*

In the space-charge-grid tube the outer grid is the control electrode and the inner grid is maintained at a fixed potential. The purpose of the inner grid is to reduce the effect of the space-charge around the filament and thereby reduce the plate resistance of the tube. The space-charge-grid tube performs the same functions as ordinary three-electrode tubes, but in general has higher mutual conductance than a three-electrode tube of similar design.

Several double function tubes and circuits are described in which both grids act as control-electrodes or in which one grid acts as control electrode and the other as a combination space-charge-grid and control or output electrode. These circuits are sometimes useful but are subject to certain definite limitations.

SEVERAL forms of vacuum tubes have been made with four electrodes, but the one which perhaps is the best known and of the greatest present interest is that containing a cathode, two grids, and an anode. The purpose of this paper is to discuss briefly some of the important characteristics of the two-grid tube and to show some of the relations between the tube constants and the performance of the tube in typical circuits.

The uses and types of two-grid tubes may be divided into three general classes depending upon the functions of the two grids; and to a considerable extent the use to which the tube is to be put influences the details of the design. These classes include the "screen-grid" tube, the "space-charge-grid" tube, and various "double function" and miscellaneous tubes and uses.

* Original Manuscript Received by the Institute, December 14, 1927.
* Presented at the Annual Institute Convention, January 11, 1928.

THE SCREEN-GRID TUBE†

The earliest work on the screen-grid tube[1,2,3,4] was done by W. Schottky in Germany. A. W. Hull[5,6] has further developed the tube by providing more complete shielding and present commercial examples of this tube, Radiotron UX-222, and Cunningham Type CX-322 as well as the larger transmitting tubes are the outgrowth of Dr. Hull's work.

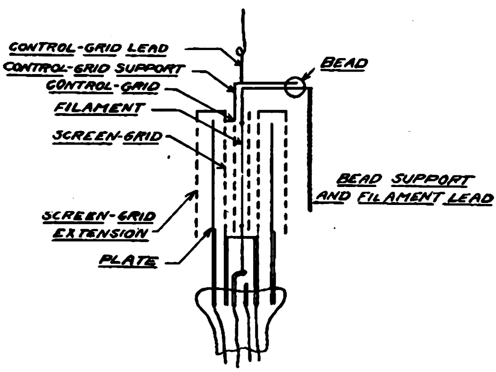

Fig. 1

Figs. 1 and 2 show the arrangement of electrodes in a typical screen-grid tube (UX-222). The inner grid is the control electrode and is similar to the grid of a three-electrode tube. It is supported

† (Author's note:—A number of names for this type of tube have appeared recently in various publications. Some of these are "shield-grid," "shielded-grid," "screened-grid," "shielded-plate," "screened-plate." These terms are obviously not all synonymous although they all have some technical justification. The last four have the disadvantage of not indicating directly that the screening or shielding is accomplished by the use of a fourth electrode. Also, there is some question as to whether the grid or the plate is screened since the screen is interposed between the two. "Screen-grid" and "shield-grid" refer directly to the fourth electrode, which is the distinguishing feature of the tube. Of these two, the name "screen-grid" has the advantage of implying both the form and function of the fourth electrode, while "shield-grid" gives the idea of a solid metal plate such as commonly is used for shielding of circuits, and in fact such as is often used around the outside of the tube. "Screen-grid" is in keeping with the idea of allowing the electrons to pass through but tending to obstruct the electrostatic field. "Screen-grid" furthermore can, without confusion or misunderstanding, be shortened to "screen" in speaking of "screen voltage" or "screen current," and the name of the tube shortened to "screen tube.")

1 Schottky, *Archiv. fur Elektrotechnik*, Vol. 8, p. 299, Dec. 1919.

2 Schottky, U.S. Patent No. 1537708, filed Aug. 27, 1919, issued May 18, 1925.

3 Barkhausen, *Jahrbuch der draht. Tel. u. Tel.* Vol. 14 p. 43, 1919.

4 Howe, *Rad. Review*, Vol. II, No. 7 pp. 337–340 July 1921.

5 Hull and Williams, *Phys. Rev.* Vol. 27, No. 4 pp. 432–438 Apr. 1926.

6 Hull, *Phys. Rev.* Vol. 27, No. 4, pp. 439–454, Apr. 1926.

entirely from the upper end of the tube and the lead is brought through a seal in the top of the bulb to a metal terminal cap. The outer grid is the screen, and it is supported on the main stem as well as by the glass bead above the plate. The plate is supported on the stem only. The screen-grid and its extension on the outside of the plate serve to shield the control-grid from the plate. Since this shield must take the form of a screen in order to allow electrons to pass through it, the shielding is not absolutely perfect, but can with a practical design be made sufficiently good to reduce the grid-plate capacity to a small fraction of the capacity in a three-

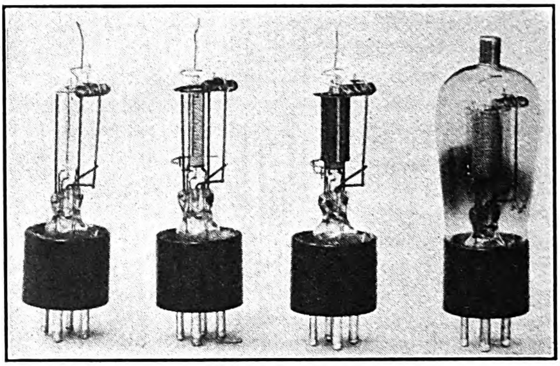

Fig. 2—UX-222

electrode tube. For example, the UX-222 has an effective capacity of about 0.02 $\mu\mu$f. Usually a metal cap should be placed around the outside of the bulb and connected to the filament or screen in order to prevent any stray coupling around the outside of the electrodes. Tinfoil wrapped on the bulb and grounded is equally effective.

Fig. 3 shows the simple circuit connections for a screen-grid tube. The screen-grid is connected directly to a positive voltage somewhat less than the plate voltage. The input circuit is connected to the inner grid in the usual manner. In a typical radio-frequency amplifier circuit the plate supply voltage may be 135 volts and the screen voltage 45 volts. Electrons leaving the cathode and passing through the control grid are accelerated toward the screen and most of them pass through it and are then drawn to the plate since there is negligible space charge between screen and plate. Some of the electrons are, of course, caught by the screen and perform no useful function, but this loss need not

be over 25 percent of the total current and is often less than 10 percent. The field at the cathode is not influenced appreciably by the plate voltage on account of this shielding of the screen and is determined by the voltages of the control-grid and the screen.

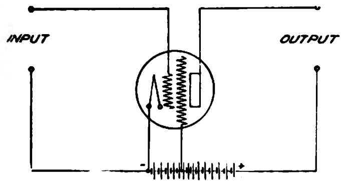

Fig. 3

Also, the amount of control of the grid voltage on the total current, i.e., the mutual conductance, is determined largely by these same voltages. The plate then acts as a collector of the electrons which have passed through the screen and, because of the electrostatic shielding between the plate and the other electrodes, the plate

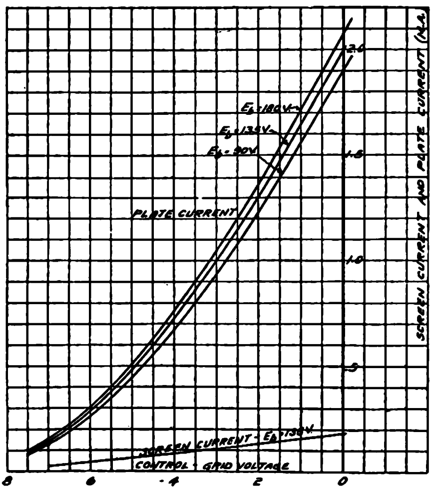

Fig. 4—UX-222 Mutual Characteristics, Screen-Grid Connection. $E_{c2}=45$ **volts.**

voltage has very little effect on the plate current. The internal plate resistance is accordingly very high, in typical cases being 500,000 to 1,000,000 ohms. This is an important advantage since, as will be explained later, the effective amplification factor of the tube is increased at the same time as the plate resistance.

Characteristics of Screen-Grid Receiving Tubes (UX-222)

Typical static characteristics of a screen-grid receiving tube are shown in Figs. 4, 5, and 6.‡ The grid-voltage plate-current characteristics of Fig. 4 are similar to ordinary three-electrode tube curves except that as the plate voltage is changed, the position of the curve is only slightly shifted. This is merely another indication of the high plate resistance. The screen current curves show the amount of current which is lost in the screen circuit. In Fig. 5 that part of the curves below the screen voltage (45 in this case)

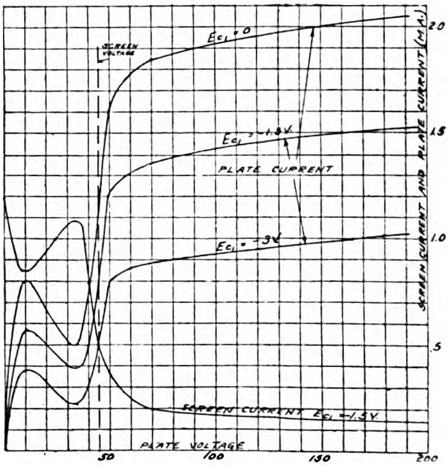

Fig. 5—UX-222 Mutual Characteristics, Screen-Grid Connection. $Ec_2=45$ **volts.**

is of little practical interest since the tube is ordinarily used at plate voltages considerably higher than the screen voltage. Secondary emission from the plate plays a considerable part in the shape of this section of the curve. Above the screen voltage the plate current curves flatten out and the plate conductance becomes only a few micromhos. If there were perfect shielding and no secondary emission from the screen the curves would be perfectly flat. Fig. 6 shows the effect of the screen voltage on its own current and the plate current. The plate current curves are very similar to the plate characteristics of a three-electrode tube.

These curves illustrate the fact that the plate current and the

‡ (Author's note:—In these figures and others, the subscripts c_1 and c_2 have been used to denote the first and second grids respectively, regardless of their functions. For example Ec_2 refers to the voltage of the outer grid regardless of whether it is being used as a control-grid or as a screen-grid.)

mutual conductance of a given tube are dependent almost entirely upon the voltages of the two grids. In fact these quantities are almost the same as they would be in a three-electrode tube having the same filament and control-grid, but with a plate at the location of the screen and at a potential equal to that of the screen. (In the three-electrode tube, the plate current would be equivalent to the sum of screen and plate current in the screen-grid tube.) The most important characteristic of the tube is the mutual conductance and the amplification factor is not necessarily a

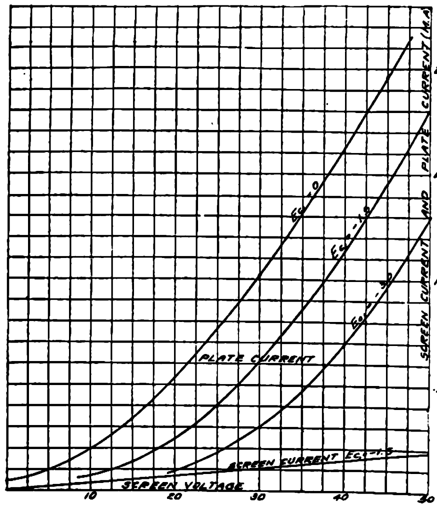

Fig. 6—UX-222 Screen Characteristics. $Eb = 135$ volts.

direct function of the geometry of the tube. If there were no secondary emission from the screen, the amplification factor would be dependent upon the diameters of the grids and plate and the mesh of the grids,[7] but even a small amount of secondary emission upsets the usual relations. The effect of this emission is to raise the mutual conductance and to lower the plate resistance since more electrons reach the plate. In the UX-222, the over-all effect is a slight increase in amplification under practical conditions. The mutual conductance of the UX-222 at 135 volts on the plate, 45 volts on the screen, and -1.5 volts bias on the control grid is about 0.35 m.a. per volt and the plate resistance averages about 850,000 ohms. This gives an amplification factor of nearly

[7] Schirmann, *Archiv. fur Elektrotechnik*, Vol. 8 p. 441, March 1920.

300. It should be remembered, however, that while this factor enters into the circuit calculation in the same way as for the three-electrode tube, it is not a constant but varies considerably with the electrode voltages as shown by Fig. 7.

There is a noticeable difference between these amplification characteristics of the screen-grid tube and those of a three-electrode tube having a very high amplification factor. It is, of course,

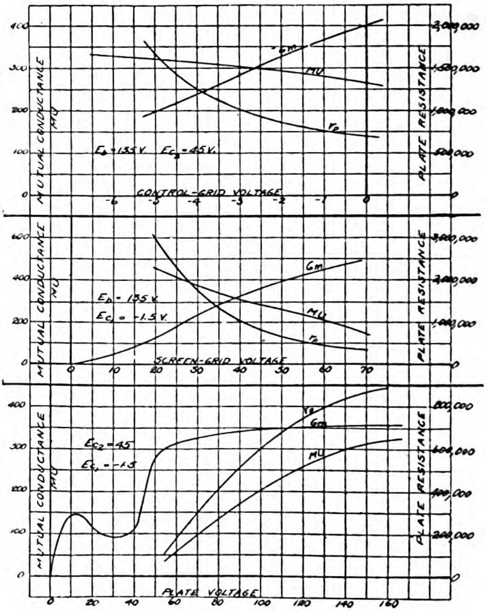

Fig. 7—UX-222 Characteristics, Screen-Grid Connection.

possible to make a three-electrode tube with an amplification factor of 200 or more by using a very close grid mesh or increasing the distance from grid to plate. However, in doing this, the mutual conductance would be reduced to so low a value that the tube would be of little practical use. In addition, the three-electrode tube would have the usual feed-back capacity.

It is perhaps self-evident that the screen-grid tube is particularly well suited to use as a radio-frequency amplifier. The absence of appreciable feed-back capacity makes it possible to use tuned coupling circuits without danger of oscillation provided, of course, that the circuits themselves are properly shielded from each other. At the same time the high impedance obtained by resonance is

desirable in order to utilize the high voltage amplification of the tube.

Some examples of the obtainable amplification may be of interest. It will be assumed that the plate resistance of the tube is 850,000 ohms, and the mutual conductance 350 micromhos, giving an amplification factor of practically 300. If the load circuit is tuned to resonance, the impedance becomes a pure resistance and in the broadcast frequency range a well designed load circuit will have an impedance of about 100,000 to 200,000 ohms.

Fig. 8—UX-222 Voltage Amplification with Resonant Loads. Screen-Grid Connection—$Ec_1=-1.5$; $Ec_2=45$; $Eb=135$.

The usual expression for voltage amplification is

$$A_v=\frac{\mu R_p}{r_p+R_p} \tag{1}$$

where R_p = load resistance

r_p = plate resistance

μ = voltage amplification factor.

However, it has already been pointed out that the mutual conductance of the screen-grid tube is more of a fundamental constant than the amplification factor and the physical significance of this is perhaps better shown by writing Eq. (1) in the form

$$A_v=g_m\frac{R_p r_p}{R_p+r_p} \tag{2}$$

That is, the voltage amplification is equal to the mutual conductance multiplied by the resistance of the load and the internal resistance in parallel.

Assuming the load to be 100,000 ohms, the voltage amplification is 31.3. For a load of 200,000 ohms this is increased to 56.5.

Fig. 8 shows the relation between load resistance and amplification for loads up to 400,000 ohms.

The use of the screen-grid tube is, of course, not limited to radio-frequency amplification, but it must always be remembered

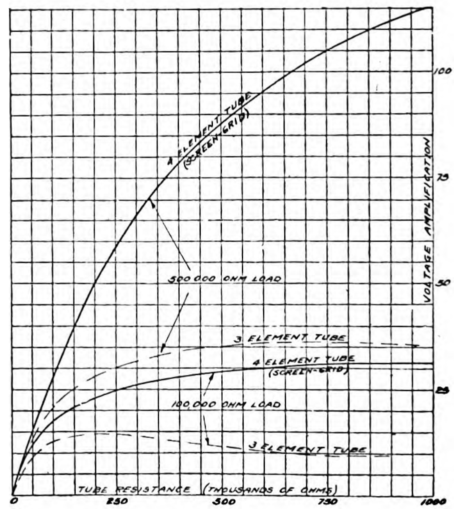

Fig. 9—Variation of Voltage Amplification with Plate Resistance.

that the voltage amplification depends almost entirely upon the load impedance, hence the over-all frequency characteristic of tube and load is very different from what would be obtained with a three-electrode tube having low plate resistance. One way of explaining this difference between the screen-grid tube and the ordinary three-electrode tube is to say that in the three-electrode tube circuit where the load impedance is usually higher than the plate resistance, the current through the load is determined more by the load impedance than by the plate resistance. In the screen-grid tube, the plate resistance is almost invariably higher than the load impedance and the current is determined mostly by the plate resistance instead of the load impedance. In the three-electrode tube it is often considered desirable to have a low plate

resistance, but in the screen-grid tube it is made as high as possible. This is simply due to what has already been mentioned, that in the three-electrode tube, an increase in plate resistance is always accompanied by a decrease in mutual conductance. Hence, for any given value of load resistance there is an optimum tube resistance for maximum amplification assuming a definite electrode structure. In the screen-grid tube an increase in plate resistance by improved screening has an inappreciable effect on mutual conductance with the result that *for any value of load resistance* the higher the plate resistance of the tube, the higher will be the voltage amplification. Fig. 9 shows the variation of voltage amplification with plate resistance for a typical three-electrode tube and a screen-grid tube. The curves for the three-electrode tube always go through a maximum at a finite value of plate resistance while the curves for the screen-grid tube rise indefinitely and approach a limit equal to $g_m R_p$.

Dr. Hull has called attention to the fact that with infinite plate resistance the mutual conductance becomes the only parameter of the tube. In the amplification equation already given,

$$A_v = g_m \frac{R_p r_p}{R_p + r_p},$$

if g_m is constant A_v increases as r_p increases until in the limit with r_p infinite, $A_v = g_m R_p$. The current in the output circuit per volt input is equal to g_m regardless of the load impedance. It is possible by means of secondary emission effects to operate a properly designed screen-grid tube under conditions which will give an infinite resistance over a limited range or even to make the resistance negative. Such secondary emission effects comprise an entire field of study in themselves and no attempt will be made to discuss them in this paper.

Screen-Grid Transmitting Tubes

In radio-frequency power amplifier circuits, the problem of preventing self-oscillation of the amplifier tube is much the same as in receiving circuits. With three-electrode tubes, it is usually necessary to employ some means for neutralizing the internal capacity of the tube to prevent oscillation when the plate and grid circuits are tuned to nearly the same frequency. As in receiving circuits the screen-grid provides a means of doing away

with the necessity for neutralization, and screen-grid transmitting tubes with output ratings up to 750 watts are being used successfully. Two of these are shown in Figs. 10 and 11. Such tubes are used in much the same way as the three-electrode power amplifier tubes except that the neutralizing circuits are of course omitted. The grid and plate circuits must be shielded from each other to prevent feedback since the shielding in the tube eliminates only the feedback through the tube itself. The elimination of the neutral-

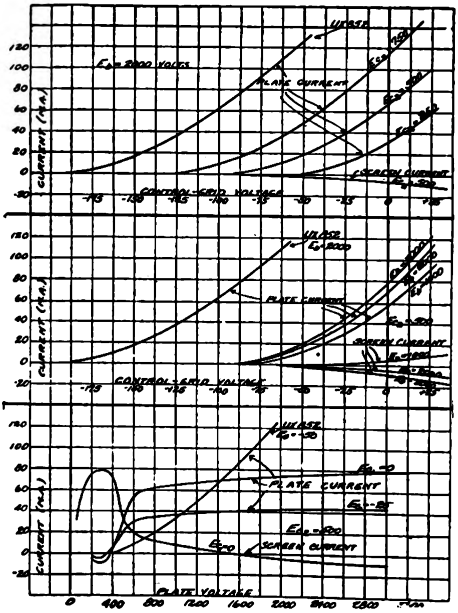

Fig. 12—Characteristics of 75-Watt Screen-Grid Tube.

izing circuits, of course, simplifies the transmitter design and, particularly in the case of transmitters designed to work on a number of different frequencies, often results in a reduction in the number of controls.

The design of the two tubes illustrated is unique in one respect in that they have been developed from two regular three-electrode transmitting tubes by the addition of the screen-grids. The smaller tube has an output rating of 75 watts and is similar to the UX-852 Radiotron while the larger tube has been derived from a 750-watt short wave transmitting tube. In each case the filament and plate and in fact practically the whole electrode structure are the same

in the four-electrode as in the three-electrode tube, the screen being mounted on the regular filament and grid stems.

Static characteristics of the 75-watt tube are shown in Fig. 12 and for comparison some of the curves for the UX-852 have been drawn in also. Similar curves for the 750-watt screen-grid tube are shown in Fig. 13.

The effect of the screen-grid on the feed-back capacity is shown by a comparison of the control-grid to plate capacities of the three

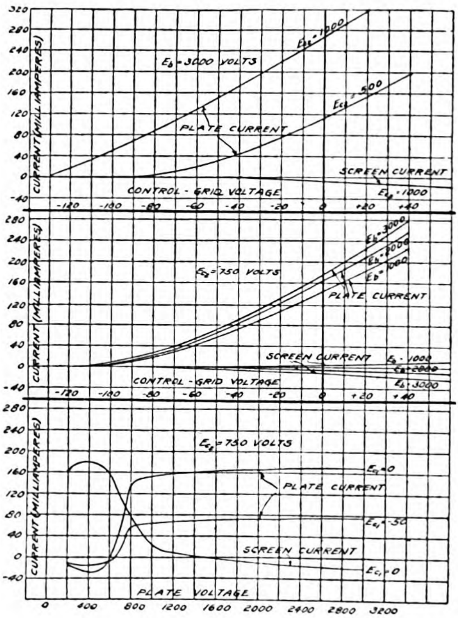

Fig. 13—Characteristics of 750-Watt Screen-Grid Tube.

and four-electrode tubes. In the UX-852 the grid-plate capacity is about 2.5 mmf., which is relatively low for a three-electrode tube since it is designed for short wave use. The screen reduces this to 0.05 mmf. In the 750-watt tubes the screen reduces the capacity from 3.9 mmf. to 0.05 mmf.

The Space-Charge-Grid Tube

The early work on this form of tube was done by Langmuir[8] in this country, and by Schottky[1] and Barkhausen[3] in Germany.

[8] Langmuir. U. S. Patent No. 1558437 filed Oct. 29, 1913, issued Jan. 1, 1924.

The tube takes it name from the fact that the inner grid is used in such a way as to reduce the space-charge around the cathode and thereby reduce the internal plate resistance of the tube. This is done without any decrease in amplification factor, hence, there is an increase in mutual conductance corresponding to the decrease in plate resistance.

The inner grid or space-charge-grid, in its simplest connection, is kept at a fixed positive potential with respect to the filament.

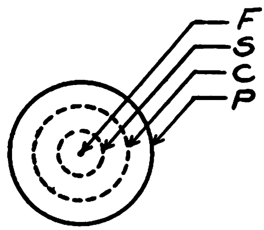

Fig. 14

The outer grid is the control electrode and performs exactly the same function as in a three-electrode tube. The arrangement of the two grids is therefore just opposite to that of the screen-grid tube.

The reduction in space charge is the result of the formation of a virtual cathode in the region between the two grids and very close to the outer one. Fig. 14 illustrates a cross section of the tube

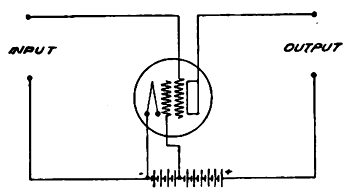

Fig. 15

and Fig. 15 shows the simple circuit connections. The positive voltage on the space-charge-grid is usually about $\frac{1}{4}$ to $\frac{1}{2}$ the plate voltage. Electrons leaving the cathode F, (Fig. 14) are accelerated by the force exerted on them by the first grid S. Some of them strike S and comprise part of the current flowing in the space-charge-grid circuit. Others pass through the meshes of the grid and since they are now travelling against the electrostatic force they come to rest in a region close to the control-grid C. Some are now drawn

back to *S*, but since others take their place there is a continuous supply of electrons which are at rest or nearly so in the region close to the control grid. The effect is much the same as if an actual cathode had been placed very close to the outer grid—much closer than could be done by mechanical means. In this way the plate resistance is reduced just as it would be in a three-electrode tube if the grid were placed very close to the filament. Of course in

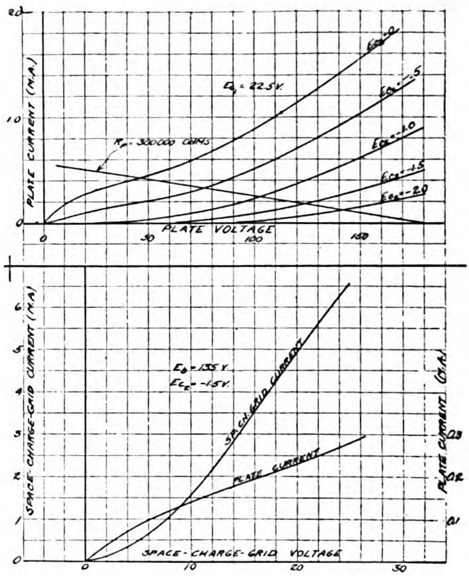

Fig. 16—UX-222 Characteristics, Space-Charge-Grid Connection.

designing a space-charge-grid tube the amplification factor and plate resistance may be made high or low as desired just as in a three-electrode tube. However, the greatest gain in mutual conductance, as compared with a three-electrode tube having the same amplification factor, occurs in a tube with a medium or high amplification factor.

Characteristics of the UX-222 as a Space-Charge-Grid Tube

The UX-222 is primarily designed to be used as a screen-grid amplifier but by proper connections to the grids it may be used as a space-charge-grid tube. Since the outer grid has a very close mesh it is obvious that the tube will have a relatively high amplification factor (compared with common three-electrode tubes)

when used in this way although not nearly so high as when used as a screen-grid tube. Under usual operating conditions the amplification factor is about 60 to 80. Naturally, the plate resistance will be high also, and the tube is therefore most useful with high impedance coupling circuits such as in resistance-coupled amplifiers.

Figs. 16 and 17 show the static characteristics of the tube with space-charge-grid connection. The plate current curves have the same general form as those of a three-electrode tube. The current to the space-charge-grid, Ic_1, is higher than the plate current at negative values of control grid voltage Ec_2, but as Ec_2 is made

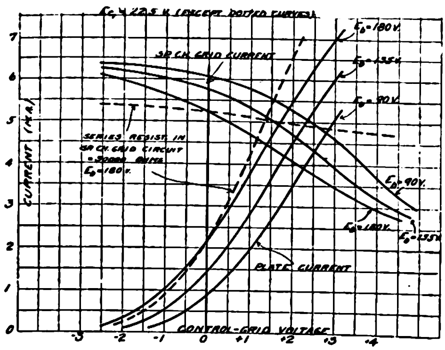

Fig. 17—UX-222 Mutual Characteristics, Space-Charge-Grid Connections, $Ec_1=22.5$ v. (except dotted curves).

more and more positive, Ic_1 decreases and eventually becomes less than the plate current. This is explained by the fact that at low values of plate current most of the electrons which pass through the space-charge-grid eventually return to it and comprise a relatively high current, but as the control-grid voltage is made more positive the plate takes more and more electrons from the space between the two grids and fewer are left to return to the inner grid. If the control-grid is made sufficiently positive, all of the electrons passing through the inner grid go on to the plate (or outer grid) and the inner grid current reaches its minimum value.

The amplification factor, plate resistance and mutual conductance of the UX-222 used as a space-charge-grid tube are shown in Fig. 18.

Using this tube in a resistance-coupled audio amplifier with a coupling resistance of 300,000 ohms, grid bias of -0.75 volt, and 180 volts plate supply voltage the actual plate voltage at the tube is 90 volts (See load line on Fig. 16). The amplification factor is

approximately 80 and plate resistance 160,000 ohms. The increase in mutual conductance due to the space-charge-grid is very apparent here. A three-electrode tube of the same general size with an amplification factor of 80 and worked at a plate voltage of 90 would have a mutual conductance of about 125 or only one-fourth as much as the space-charge-grid tube.

One of the limitations in the use of tubes with very high amplification factor in resistance-coupled amplifier circuits is that

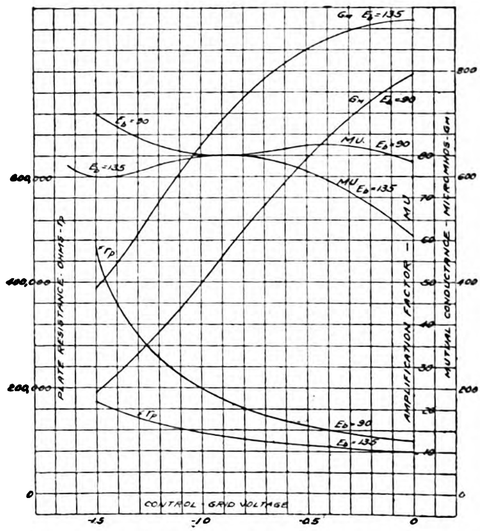

Fig. 18—UX-222 Characteristics with Space-Charge-Grid Connection. $E_{c1} = +22.5$ volts.

their effective input capacity is apt to be very high. This results in a falling off of amplification at the higher audio frequencies. The higher the amplification of the stage, the more serious this effect becomes because the input capacity is

$$Cg = Cgf + Cpg(A_v + 1). \tag{3}$$

where Cgf = direct grid-filament capacity
Cpg = direct plate-grid capacity
A_v = voltage amplification of stage.

This same limitation applies to the space-charge-grid tube having high amplification factor. As will be shown later the screen-grid tube used with resistance coupling does not have this high input capacity although it does have another sort of limitation.

Before leaving the subject of the space-charge-grid characteristics it may be well to repeat that all that has been said about the high amplification factor and plate resistance applies only to the UX-222 and not to all space-charge-grid tubes. These tubes can be made with any desired amplification factor, the same as a three-electrode tube. Tubes with a factor of 6 to 10 have been used quite extensively in Europe for a number of years. European practice has been to design and use the tubes in such a way as to permit very low plate voltages, of the order of 6 to 20 volts. Some work has been done in England with no plate voltage except the

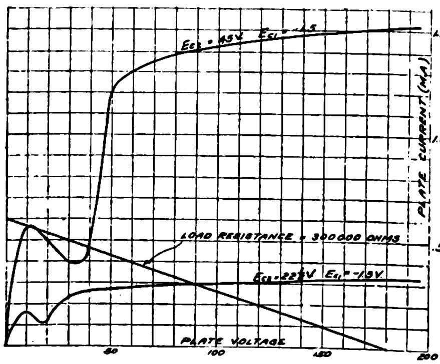

Fig. 19—UX-222 Characteristics with Resistance Coupling.

filament drop, but the performance in this case is not as good as when at least 6 or 8 volts are used. It might seem that a space-charge-grid could be used to raise the mutual conductance of a power amplifier tube but this has not yet been done very successfully. A power amplifier tube must ordinarily work with high bias and high a-c. input voltage. In a tube containing a space-charge-grid the curvature of the plate-voltage plate-current curves at high negative grid voltages is such that the distortion is abnormally high. Hence, the increase in power output is not as great as might be expected from the large increase in mutual conductance which occurs at lower negative grid bias and with small amplitudes.

The Screen-Grid Tube as an Audio Amplifier

An examination of the characteristics of the screen grid tube indicates that while the tube cannot be used as an audio amplifier in the ordinary way with transformer coupling on account of its

high plate resistance, it may be used with resistance coupling. It has already been pointed out that when the UX-222 is used as a space-charge-grid tube with resistance coupling the voltage amplification is high but falls off at the higher frequencies. The screen-grid tube has a grid-plate capacity of only 0.02 mmf. hence, even when multiplied by a voltage amplification of 40 is only a small part of the total input capacity, which is therefore practically equal to the grid-filament capacity itself. However, the amplification given by the screen-grid tube is not as great as might be expected. This is explained by reference to Fig. 19. If it is assumed that the supply voltage is 180, a series load will give a load characteristic as shown by the straight line passing through the 180-volt point and having a slope equal to the reciprocal of the resistance.

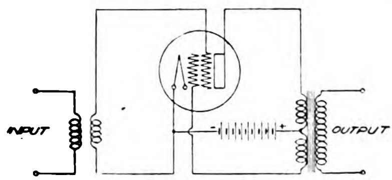

Fig. 20

This intersects the plate current curves for $Ec_1 = 0$ and $Ec_1 = -1.5$ at points far outside the normal operating range, and off the flat part of the plate characteristic. There are two ways of correcting this condition—the screen voltage may be lowered or the control-grid voltage made more negative. Either procedure will lower the mutual conductance but this cannot be avoided. Referring again to Fig. 19, the lower curve represents proper operating conditions and here the voltage amplification for small swings is about 40.

Double Function and Miscellaneous Uses

Most of the four-electrode tube applications of this class may be divided into two groups—first, those in which one of the grids is made to serve as a space-charge-grid and at the same time as a control electrode or as an output electrode, and second, those in which the two grids are control electrodes only. Numerous varieties of tubes and circuits have been proposed and no attempt will be made to describe all of these in this paper. Only a few will be discussed in order to illustrate some of the general principles involved.

An examination of the static characteristics of a space-charge-grid tube immediately suggests the possibility of making use of the variation of space-charge-grid current with control-grid voltage, and a large number of schemes for doing this have been proposed. The variations in grid current are of course opposite to the variations in plate current and it is a simple matter to combine the output of both circuits. For example, an output transformer with a tapped primary connected as shown in Fig. 20 will combine the two outputs in correct phase relation.[9] A similar result may be obtained with resistance coupling as in Fig. 21. In the latter case there is an added advantage in that no direct current need flow through the load if the resistances are properly balanced. This is sometimes important when the tube is used for measurement

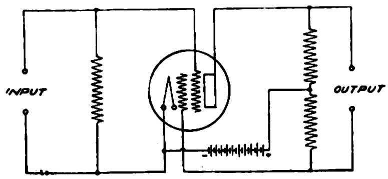

Fig. 21

purposes and also permits the amplification of d-c. voltages. Another sort of push-pull arrangement for radio-frequency amplification has been proposed in France[10] in which the space-charge-grid performs its usual function and at the same time acts as a neutralizing condenser. The adjustment of this circuit is apt to be somewhat critical because the alternating voltage at the space-charge-grid which is used for neutralizing is partially dependent upon the electron emission of the filament. Hence, a close adjustment of filament temperature is required.

All of these schemes have been shown with common battery for plate and space-charge-grid. If, as is usually the case, the best space-charge-grid voltage is less than the plate voltage a series resistance may be used to reduce this voltage to the right value, which introduces another effect of some interest. The circuit arrangement is shown in Fig. 22. When the control-grid voltage is varied the current flowing through the resistance varies in a direction opposite to that of the plate current, that is, when the plate current rises the space-charge-grid current falls and conse-

[9] VanderBijl, U.S. Patent No. 1479779, filed July 20, 1920, issued Jan. 1, 1924.

quently causes the voltage between space-charge-grid and filament to rise. This voltage exerts a certain amount of control upon plate current and although less in magnitude, it is the same in direction as the control of the control-grid voltage, hence, the rise in space-charge-grid voltage aids the original change in control-grid voltage. The amplification is therefore somewhat greater with the resistance in the inner grid circuit even though the mean space-charge-grid voltage is the same. The dotted curves in Fig. 17 show the effect of using this resistance. The space-charge-grid current varies only slightly with varying control-grid voltage while the plate current changes more rapidly than without the resistance.

Another method of making use of the variation in space-charge-grid current is to couple this grid circuit back to the input and produce regeneration. Obviously, the reverse effect may be

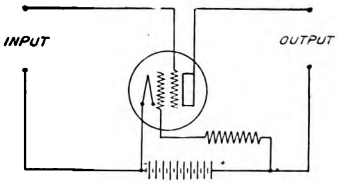

Fig. 22

used to counteract any undesirable regeneration which may already exist due to coupling from the plate circuit to the input.

At first thought it might seem that by virtue of the variation of space-charge-grid current this type of tube should give the same results as two three-electrode tubes in any of the balanced circuits such as the push-pull circuit. This is not the case, however, because the a-c. power which may be drawn from the space-charge-grid circuit is strictly limited. That is, while the control-grid may produce large current changes in the space-charge-grid circuit when the latter contains no external impedance, the current changes become much smaller when a load circuit is connected. Otherwise stated, the voltage amplification factor of the control grid with respect to the space-charge-grid $\left(-\frac{\partial e_{g_1}}{\partial e_{g_2}}\right)_{i_{g_1}\text{ const.}}$ and the internal output resistance of the space-charge-grid circuit $\left(\frac{\partial e_{g_1}}{\partial i_{g_1}}\right)$ are low even though the mutual conductance $\left(\frac{\partial i_{g_1}}{\partial e_{g_2}}\right)$ is relatively high.

In the second group of double function circuits in which both grids act as control electrodes alone may be mentioned the many reflex circuits, regenerative and super-regenerative circuits, frequency changers, etc., which have appeared in foreign publications.[10–18] The number of combinations of electrode functions is almost unlimited and many of them have a certain degree of usefulness although few have come into common use. This is probably due to the two limitations to which practically all of this group of double function schemes are subject. In order to act as control electrodes, the two grids must be operated at a mean potential of zero or slightly negative so as not to load the input circuit. With the two grids between filament and plate and at this low potential the plate resistance is relatively high and the control of either grid is relatively low, and less than would be obtained in a three-electrode tube under corresponding conditions. The other limitation applies to circuits in which both grids have radio-frequency voltages impressed upon them, as for example, one grid excited by antenna and the other coupled to the plate circuit for regeneration. This limitation is the electrostatic capacity between the two grids which ordinarily makes it impossible to vary the potential of one grid without also affecting the other.

One other type of double function circuit uses the plate as one of the control electrodes and the output is taken from one of the grids.[19]

Conclusion

Particular attention should be called to the foreign publications on four-electrode tubes. Those which have been given as references represent only a small part of all that have appeared and they have been chosen merely as representative of the work which has been going on for a number of years. Strangely enough the screen-grid tube although developed from Schottky's early work has been mentioned very little in the foreign literature until very recently,

[10] Barthelemy, *L'Onde Electrique*, No. 64, pp. 152–160 Apr. 1927.
[11] Scott–Taggart, *The Electrician*, Jan. 21, 1921 pp. 97–98.
[12] De Mare. *La T. S. F. Moderne*, No. 28, pp. 494–500 Oct. 1922.
[13] deVoogt, *Radio Nieuws*, 4, p. 99, Apr. 1921.
[14] Corver, *Radio Nieuws*, 4, p. 104, Apr. 1921.
[15] Nozieres et Giroud, *L'Onde Electrique*, pp. 583–590 Dec. 1924.
[16] Chauvierre, *QST Fr. et Radioelectricite*, Vol. 8, pp. 8–16, Apr. 1927.
[17] Mittelman, *Der Radio Amateur*, Vol. 4, No. 29 pp. 571–573, No. 30, pp. 595–597, No. 31 pp. 605–607, No. 32 pp. 629–632 1926.
[18] Decaux, *L'Onde Electrique*. No. 61, pp. 1–18 Jan. 1927.
[19] Donisthorpe, Proc. I.R.E. Vol. 12, No. 4, pp. 411–421 Aug. 1924.

apparently much more work having been done on space-charge-grid tubes and on double-function circuits.†

The degree of general usefulness of the various four-electrode tubes and circuits may perhaps be expressed by saying that the space-charge-grid tube performs the same kind of functions as the three-electrode tube but at lower plate voltages or with somewhat higher amplification; the double function circuits while often very interesting in themselves, accomplish with one tube what can often be done almost as simply, as effectively, and sometimes less expensively with two three-electrode tubes; but the screen-grid tube not only permits a degree of radio-frequency amplification much greater than can be obtained with a three-electrode tube but also eliminates the feedback which is so often an unwanted function of the three-electrode tube.

In conclusion, the writer wishes to thank Messrs. O. W. Pike, A. C. Rockwood, and B. J. Thompson for their assistance in the preparation of the data given on the screen-grid tubes.

† (Author's note:—Since the first draft of this paper was written, screen-grid receiving tubes have been placed on the market in England and several excellent articles on the subject have appeared in English publications. See references 20, 21, and 22 below.

[20] McLachlan, *Wireless World*, Vol. XXI pp. 260–263 Aug. 31, 1927 and pp. 307–310 September 7, 1927.

[21] McLachlan, *Exp. Wireless and Wireless Eng.*, Vol. IV pp. 597–600 October 1927.

[22] Beatty, *Exp. Wireless and Wireless Eng.*, Vol. IV pp. 619–625 October 1927

THE INVERTED VACUUM TUBE, A VOLTAGE-REDUCING POWER AMPLIFIER*

Frederick Emmons Terman

(Stanford University, California)

Summary—*By interchanging the functions of the grid and plate of the usual vacuum tube a voltage-reducing power amplifier is obtained. The usual vacuum tube acts as a voltage-increasing power amplifier.*

The static curves of the inverted vacuum tube are similar in form to the corresponding curves of the ordinary vacuum tube, and the theory of the inverted vacuum tube is analogous in all respects to the usual vacuum-tube theory, the only difference being reduction instead of amplification of voltage.

It is relatively simple to construct an inverted vacuum tube with wide clearances between plate and the rest of the tube, so that potentials of hundreds of thousands of volts can be applied to the plate, while the effect of this high voltage stepped down in almost any desired ratio is obtained in a low-potential circuit.

Introduction

THE inverted vacuum tube is a three-element vacuum tube that has been made to operate as a voltage step-down device by interchanging the grid and plate functions in the ordinary tube circuit. The basic circuit of the inverted vacuum tube is shown in Fig. 1. The essential features of this circuit are: (1) the grid is operated at a positive potential and so draws considerable

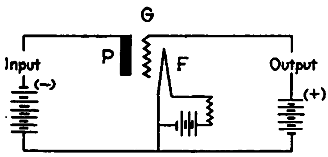

Fig. 1—Circuit of the Inverted Vacuum Tube.

current; (2) the plate is operated at a negative potential and so draws no current; and (3) the input circuit is the plate circuit, while the output circuit is the grid circuit. The inverted vacuum tube is merely an ordinary vacuum tube with the grid doing what the plate usually does, and with the plate carrying on the functions usually performed by the grid.

The inverted vacuum tube acts as a voltage reducer, or voltage step-down device, because a voltage applied at the input, *i.e.*, to the plate, has the same effect on the grid current as though a similar but much smaller voltage had instead been added to the grid battery. If the vacuum tube used has a voltage amplification

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* Presented at New York Institute Meeting, February 1, 1928.

factor of μ, the step-down ratio will be approximately μ. The inverted vacuum tube acts as a power amplifier, for with the plate negative practically no energy is consumed in the plate circuit when a voltage is applied to the input, while at the same time this input voltage controls an appreciable grid current, thus controlling considerable energy in the grid, or output, circuit.

The inverted vacuum-tube circuit was developed for the purpose of taking oscillograms of voltages without disturbing the action to be observed by drawing current to operate the oscillo-

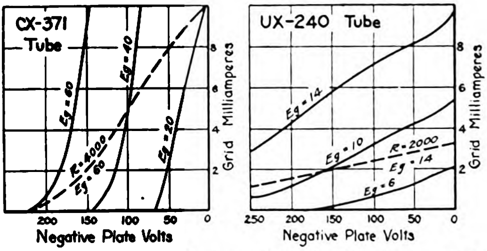

Fig. 2—Relation of Plate Voltage and Grid Current at Constant Grid Volts.

graph vibrator. The device will also function as a generator of either radio- or audio-frequency oscillations, as an amplifier of voltages, as a voltmeter, as a rectifier, etc. These applications are given detailed treatment in a later section of the paper.

Static Curves

The operation of the inverted vacuum tube can best be visualized in terms of curves giving the relation between grid current and plate and grid voltages. A complete set of such curves for two representative tubes is given in Figs. 2, 3, 4, 5, and 6. One of these tubes has a μ of about 3 (step-down ratio of approximately 3) while the other has a μ of about 30 (step-down ratio of approximately 30). As the two companion sets of curves are plotted to the same scales, and apply to tubes very similar to each other in general construction except for the value of μ, the effect of changing the step-down ratio can be readily visualized. All potentials in the figures are measured with respect to the negative filament lead.

The curves of Fig. 2 show the relation between grid current and negative plate voltage for various values of positive grid voltage. The results are seen to have a form similar to the $E_g - I_p$

(grid-voltage, plate-current) curves of the usual vacuum-tube circuit. An important feature is that each curve in Fig. 2 has a straight line portion of more or less length. Variations of plate voltage within the boundaries of this straight line portion give rise to variations in grid current that are directly proportional to the plate voltage changes. When operating on this straight line part

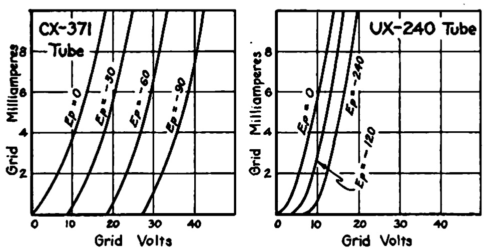

Fig. 3—Relation of Grid Current to Grid Voltage at Constant Plate Volts.

of the characteristic *the grid current variations give a distortionless representation of any a-c. or d-c. voltage applied at the input shown in Fig. 1.*

The dash lines in Fig. 2 give the characteristic when a resistance is inserted in series with the grid. The effect of this resistance is to straighten out the characteristic, and to extend greatly the range of plate voltage over which the grid current is proportional to changes in plate potential.

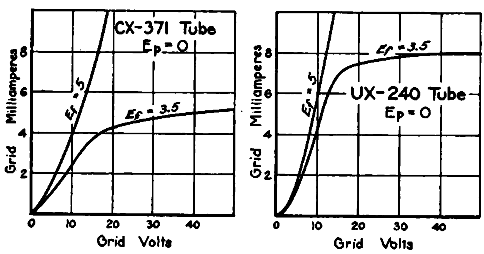

Fig. 4—Static Characteristic Illustrating Effect of Saturation

Fig. 3 gives the same information as Fig. 2 but in another form. The curves of Fig. 3 correspond to the usual $E_p - I_p$ curves of the vacuum tube, and have the same form. The effects of limited electron emission from the filament are shown in Fig. 4. It is seen that as long as the filament temperature is sufficient to emit a small surplus above the number of electrons flowing to the positive grid, the filament temperature is unimportant. Portions of the curves in Fig. 4 where the electron emission is insufficient to furnish

the necessary supply of electrons required for normal operation are flattened out as the figure indicates. The departure from normal takes place at lower and lower grid currents as the filament voltage is reduced because the electron emission drops rapidly with reduction in filament heating.

The curves in Fig. 5 are very important in the theory of the inverted vacuum tube, and will be discussed in the next section.

Fig. 6 differs from Fig. 2 only in showing the effect on the grid current of making the plate voltage positive. It is seen that as soon as the plate begins to take current, the grid circuit is robbed

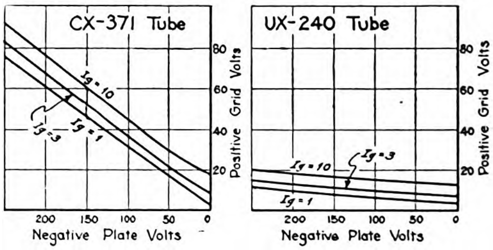

Fig. 5—Relation of Plate and Grid Voltages Required for Constant I_g.

of a large part of its current. The very sudden change of curvature in the characteristics at zero plate volts indicates that this point gives a large rectifying effect.

Theory

The theory of the inverted vacuum tube rests upon the fundamental fact that the current flowing to the positive grid depends upon the electrostatic field existing between filament and grid, and is substantially independent of how the field is produced. Both the grid and plate potentials affect the magnitude of this field. It can be shown from the theory of electrostatics that the actual field is proportional to the quantity (E_g+E_p/m), where E_g and E_p are the potentials of the grid and plate respectively, and m is a constant that is determined by the geometrical proportions of the tube and is not affected by the electrode potentials. Stating that the field intensity between grid and filament is proportional to the quantity (E_g+E_p/m) is equivalent to saying that E_g volts applied to the grid produce the same field intensity as is produced by the application of mE_g volts on the plate. The constant m is usually larger than unity, ranging commonly from 3 to 40. The constant m is larger than unity because the grid

is so constructed and located that it is in a better position to produce a high field intensity between filament and grid than is the plate.

The grid current is determined by the field intensity between filament and grid, and this field is proportional to the quantity (E_g+E_p/m). For constant grid current it follows that (E_g+E_p/m) must be constant. Fig. 5 shows a series of curves for different constant values of grid current. These curves are practically straight lines that are substantially parallel. The equation of a straight line in Fig. 5 is:

$$(E_g+E_p/m)=\text{Constant}$$

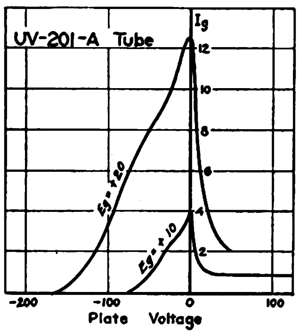

Fig. 6—Complete Static Characteristic of Inverted Vacuum tube.

where $(1/m)$ is the slope of the line. Since the lines are straight it follows that a constant grid current is obtained for all combinations of voltages that keep the quantity (E_g+E_p/m) constant. Since the factor m depends only on the slope of the lines in Fig. 5 and the lines are nearly parallel, it follows that m is a constant substantially independent of the tube voltages and current.

The lines in Fig. 5 depart a slight amount from perfect parallelism, and perfect straightness. To this extent the actual tube behavior departs from the simple theory that makes m a geometrical constant of the tube independent of electrode voltages, and that makes the grid current depend only on the value of the quantity (E_g+E_p/m).

The constant m is the same as the constant μ of the usual vacuum-tube circuit. Tests show that when the tube is used in the inverted manner, m and μ as determined from measurements are not exactly the same, m being in general slightly greater than μ. For this reason the symbol m is used for the μ of the inverted tube.

The constant m is the step-down ratio of the inverted vacuum tube. According to the theory, an increment of ΔE_p in plate potential will have an effect on the grid current equivalent to the effect of adding $\Delta E_p/m$ volts in the grid circuit. *It follows from this that the effect on the grid circuit of an input voltage e_s in Fig. 1 is exactly the same as though this input voltage had been replaced by a fictitious generator of voltage e_s/m acting in the grid or output circuit.*

The equivalent circuit of the inverted vacuum tube is given in Fig. 7. The effect of a signal voltage e_s applied to the input (plate) circuit is as though a generator having a voltage $-e_s/m$ was inserted between the grid and filament of the tube. The minus sign is present because of the direction in which this voltage is assumed to act in Fig. 7 which is the most convenient direction.

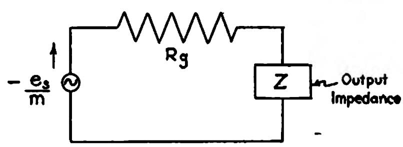

Fig. 7—Equivalent Circuit of the Inverted Vacuum Tube.

The resistance R_g is the dynamic grid resistance, *i.e.*, R_g is the resistance the grid circuit offers to an increment of voltage such as $-e_s/m$. It is the a-c. resistance of the grid circuit, and is analogous to the dynamic plate resistance R_p of the usual tube circuit. The value of R_g depends on the grid and plate-battery potentials, and can be determined by methods given in the next section. The voltage e_s that is applied to the input will generally be an a-c. voltage of magnitude E_s and frequency f. This voltage superimposed on the negative plate battery potential by being introduced at the point marked "Input" in Fig. 1, is equivalent to a voltage of amplitude $-E_s/m$ and of frequency f acting in the equivalent circuit. The current that this equivalent voltage produces depends upon the impedance of the equivalent circuit. This circuit consists of the dynamic grid resistance R_g in series with whatever impedance is inserted in the grid circuit at the point marked "Output" in Fig. 1. The grid current produced as a result of applying the a-c. input voltage of $e_s = E_s \sin \omega t$ is then

$$\text{a-c. grid current} = \frac{-(E_s/m)\sin \omega t}{R_g + Z}$$

where Z is the output impedance inserted in the grid circuit.

An important property of the inverted vacuum tube is the amount of change of grid current obtained with a given change of plate voltage, and with no load or output impedance present in the grid circuit. The constant expressing this property is the ratio dI_g/dE_p, and can be called the reflex mutual conductance of the inverted vacuum tube, and given the symbol g_r. The reflex

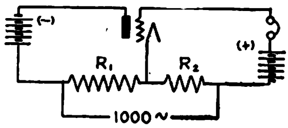

Fig. 8—Circuit for Measuring the Step-down Ratio of the Inverted Vacuum Tube.

mutual conductance of the inverted vacuum tube is analogous in all its properties to the mutual conductance g_m of the usual vacuum-tube circuit. In terms of m and R_g the reflex mutual conductance is

$$g_r = 1/mR_g$$

When there is no load impedance inserted in the grid circuit the application of a small voltage E to the plate circuit causes a change of grid current of Eg_r.

Measurement of Constants of the Inverted Vacuum Tube

The voltage step-down ratio m of the inverted vacuum tube can be measured by means of the bridge circuit shown in Fig. 8.

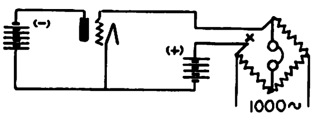

Fig. 9—Circuit for Measuring the Dynamic Grid Resistance of the Inverted Vacuum Tube.

When no sound is heard in the telephones, an elementary application of the theory of the inverted vacuum tube shows the resistances R_1 and R_2 must be such that

$$m = R_1/R_2$$

Resistances for R_2 as low as ten ohms are satisfactory. Either of the resistances may be varied to establish the balance. The actual grid voltage to which the measurement applies is the battery voltage minus the voltage drop of the grid current in flowing through the phones and R_2. To make this correction small.

low resistance phones and a moderately small value of R_2 are desirable.

The dynamic grid resistance R_g can be measured by the use of an alternating current bridge, as shown in Fig. 9. The scheme merely consists in making the grid-filament resistance the X arm of the bridge. The grid potential to which the measurements apply is the battery voltage minus the voltage drop in the bridge of the

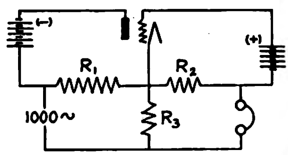

Fig. 10—Circuit for Measuring the Reflex Mutual Conductance of the Inverted Vacuum Tube.

grid current. By proper arrangement of bridge resistances the correction can be kept small.

The reflex mutual conductance g_r of the inverted vacuum tube can be measured by the circuit shown in Fig. 10. When the resistances are such as to give no sound in the phones, an elementary application of the inverted vacuum-tube theory shows that

$$g_r = R_3/R_1R_2$$

Any one of the three resistances may be varied, although R_3 is

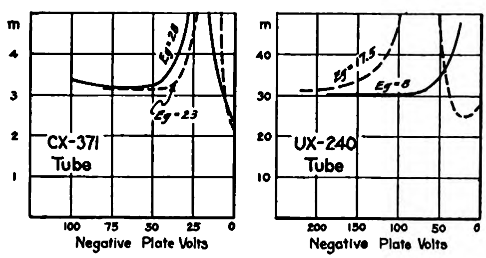

Fig. 11—Step-down Ratio of Inverted Vacuum Tube.

generally most satisfactory for the variable element. Suitable values are $R_1 = 1000$ ohms, $R_2 = 100$ ohms, and R_3 variable up to 100 ohms in tenth ohm steps. Correction for the voltage drop in R_2 due to grid current must be made to get the grid voltage from the battery potential.

In all of these measurements it is desirable to make the a-c. input voltage to the measuring equipment small. With large inputs there is some second harmonic current generated, which

masks the fundamental tone for which the bridge is to be balanced. A microphone hummer is a suitable source of audio-frequency energy for the measurements.

The constants of the inverted vacuum tube can also be derived from the static curves by methods analogous to those applied to static curves of ordinary vacuum tubes. The constants can also be determined by adding voltage increments in to the proper batteries, and reading the results on meters.

Constants of the Inverted Vacuum Tube

Results of measurements of m, and R_g for a low and a high μ tube are given in Figs. 11 and 12. Examination of Fig. 11 shows that the voltage step-down ratio m is subject to considerable fluctuations in value. Tests of a large number of tubes under

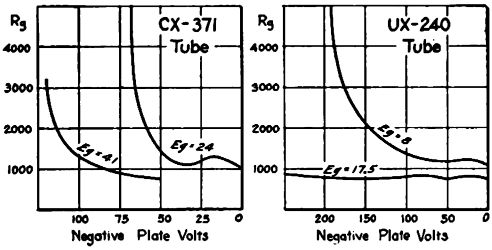

Fig. 12—Dynamic Grid Resistance of Inverted Vacuum Tube.

different conditions indicate that the step-down ratio m varies in an erratic manner when the grid current is large, and particularly when the negative plate voltage is small. For large negative potentials on the plate, and where the grid current is relatively small, the step-down ratio m is practically constant at a value approximately equal to the μ of the tube.

The dynamic grid resistance of the inverted vacuum tube is rather low, even for very low grid voltages. This is because of the proximity of the grid and filament in standard tubes. For ordinary receiving tubes, resistances in the order of 1000 ohms are found for grid potentials of around 25 volts. The dimensions of the grid do not seem to be so very important, for Fig. 12 shows that two tubes with very different grid constructions have grid resistances in the same order of magnitude.

Practical Applications of the Inverted Vacuum Tube

Oscillograph work: The inverted vacuum tube was originally developed as a means of taking voltage oscillograms in vacuum-tube circuits where the voltage was high, and where circuit

operation would be affected by even a small consumption of current in the measuring device. The inverted vacuum tube is ideally suited for this purpose. The voltage wave to be photographed is applied directly to the plate using suitable plate bias battery. By the reducing action of the tube this high voltage is transformed into a smaller potential acting in the low impedance grid circuit, developing current variations that can be registered by the oscillograph.

The best circuit for oscillograph work is that given in Fig. 13. The voltage of the plate bias battery should be such as to cause the tube to operate on a straight line portion of its characteristic. The resistance R is to straighten out the characteristic of the tube,

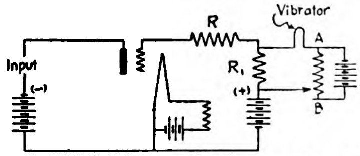

Fig. 13—Circuit for Operation of Oscillograph by Inverted Vacuum Tube.

and should be at least twice the dynamic grid resistance. The resistance R_1 and the potentiometer AB balance the normal steady grid current out of the vibrator. Values of 100 to 1000 ohms are suitable for R_1, while AB should have considerably less resistance than R_1, such as one-tenth as much. In some cases it may be satisfactory to put the vibrator directly in series with the grid circuit, and omit R_1 and the associated potentiometer.

The distortion introduced by the tube can be determined from the static curves, and the oscillograms corrected accordingly. This distortion can be minimized by making use of the straight line portions of the characteristic, and also by inserting a resistance in series with the grid battery, shown at R in Fig. 13. This resistance has the effect of straightening out the characteristic, which becomes practically a straight line if R is two or three times the dynamic grid resistance. When the resistance is used a higher grid battery potential must be employed to make up for the additional voltage drop.

The vibrator of an ordinary oscillograph requires from 50 to 100 milliamperes of current to give a good deflection. This amount of current will generally call for two 7.5-watt tubes in parallel, or a single 50-watt tube, if distortion is to be avoided in the transformation.

By reason of its practically infinite resistance, the plate circuit of the inverted vacuum tube can be safely introduced into any high-voltage low-amperage circuit. By the use of a tube with the proper step-down ratio, and with suitable insulation resistance, the high voltage to be photographed is stepped down without additional apparatus. The oscillograph is in a low potential circuit, and can accordingly be handled easily and safely. In cases where one desires to photograph a wave consisting of an alternating current potential superimposed on a d-c. potential, as in the case of the plate to filament voltage of an oscillating vacuum tube, the plate biasing battery can frequently be dispensed with.

Voltage amplification: In spite of the fact that it is fundamentally a voltage step-down device, the inverted vacuum tube

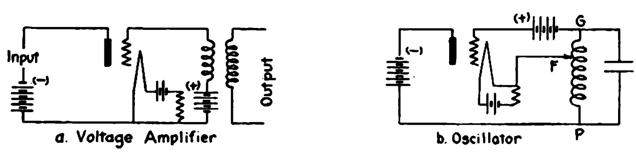

Fig. 14—Circuits for Practical Applications of the Inverted Vacuum Tube.

can function as a voltage amplifier by the use of an out-put transformer of suitable ratio. The circuit for voltage amplification is shown in Fig. 14-a. Amplification of voltage will be obtained when the step-up ratio of the transformer more than offsets the step-down ratio of the tube. High transformer ratios may be used because the dynamic grid resistance R_g that is in series with the primary is much lower than the plate resistance of ordinary amplifiers.

In an actual test using a CX-371 power tube with a step-down ratio shown in Fig. 11, and an RCA 9–1 ratio transformer, a voltage amplification between 2.6 and 2.7 was obtained for the band of frequencies from 300 to 2000 cycles. The flatness of this characteristic is due to the low value of R_g in the inverted vacuum tube.

Oscillator (*or power amplifier*): Since the inverted vacuum tube is a power amplifier it will act as an oscillator. Using the circuit shown in Fig. 14b, with +66 volts on the grid and −152 volts on the plate of a CX-371 tube, an oscillating current of 200 milliamperes at approximately 36 kilocycles was obtained in a low-resistance circuit. Because of the voltage-reducing action of

the plate on the grid circuit, it is necessary that the plate portion of the oscillating coil be many times greater than the grid portion. In the example just mentioned, best results were obtained when the part *FP* had eight times as many turns as *FG*. The oscillator requires a high negative plate potential for satisfactory operation. Inserting a condenser in series with the plate, and omitting all leak other than that furnished by the condenser dielectric will furnish this bias after a fashion, but is not so satisfactory as a battery.

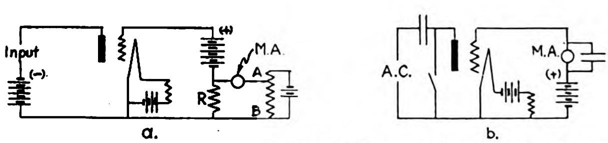

Fig. 15—Circuits for Using the Inverted Vacuum Tube as a Voltmeter.

Detection or rectification: The inverted vacuum tube can be used as a rectifier because of the non-linear properties it possesses. This rectification can be in the grid circuit (equivalent to ordinary anode rectification) or it can take place in the plate circuit with the aid of a condenser in series with the plate with a leak shunted around this condenser. Greatest sensitivity is obtained when the plate leak-condenser method is used, with little or no plate bias. The higher the resistance of the plate leak the more sensitive the detection.

The curves in Fig. 6 show a very sharp peak at the point of zero plate voltage. By operating at this point very sensitive grid circuit rectification can be obtained.

Measurement of high voltages: The inverted vacuum tube will very easily measure high d-c. voltages without drawing power from the d-c. source. The negative terminal of the unknown voltage is connected to the plate, and the magnitude of this potential is determined by the amount of grid current that is registered by a milliammeter in the grid circuit. From the static curves it is obvious that with a given grid voltage the grid current is determined by the negative plate voltage. By using the circuit shown in Fig. 15a, and making *R* rather large, and adjusting the voltage across *AB* until the milliammeter reads zero when the plate and filament are short-circuited, the milliammeter reading is practically direct reading in input volts, and is substantially independent of small variations in grid and filament battery voltages. Using a tube with a step-down ratio in the order of thousands (such a tube could

be easily made), d-c. voltages of hundreds of thousands of volts could be easily measured in a safe manner, without drawing any current from the high voltage. The measuring equipment would require a filament battery, and about 90 volts grid battery, and everything except the plate lead could be at low potential.

Alternating current voltages can be measured very conveniently by the circuit shown in Fig. 15b. The procedure is to short-circuit the plate and filament and observe the grid current. The alternating current to be measured is then applied as shown in Fig. 15b, which causes the grid current to decrease. The crest value of the a-c. wave is then the same as the value of negative d-c. voltage that would have to be applied to the plate to produce the observed reduction in grid current. The condenser C must be able to withstand twice the crest value of the voltage to be measured and should have a capacity in excess of the plate-filament tube capacity. The circuit operates by reason of the fact that C accumulates a charge whenever the plate potential becomes the least bit positive. With the application of the alternating current the condenser accumulates charge until the voltage drop across the condenser equals the crest value of the applied voltage, when no more charge accumulates. This type of a-c. voltmeter can be used for large voltages, and consumes practically no power.

Input Impedance of Plate Circuit

The plate circuit of the inverted vacuum tube has practically infinite resistance because the high negative potential at which the plate operates repels all electrons. The input resistance is not appreciably lowered by small amounts of ionization of residual gas in the tube because the positive ions are in the main produced between the filament and grid, and accordingly migrate toward the filament, not to the plate. With a negative plate voltage of 60 volts, a CX-371 tube gave a plate current of less than 2×10^{-8} amperes, corresponding to an input resistance of 3,000,000,000 ohms. This small loss apparently was due to leakage over the glass walls of the tube, for it did not change when the filament current was turned off.

The input capacity of the inverted vacuum tube is likewise small, being substantially the capacity existing between the plate as one side, and the filament and grid connected together as the other side. Since the inverted vacuum tube is a voltage reducer,

the input capacity is not appreciably increased by impedance in the output circuit, as is the case with the usual vacuum-tube amplifier.

Possibilities and Limitations of the Inverted Vacuum Tube

While the inverted vacuum tube can be used as an oscillator and voltage amplifier, it is inherently less satisfactory than the usual vacuum tube in the performance of these operations. The practical possibilities of the inverted vacuum tube appear to lie principally in oscillograph work, and in the measurement of a-c. and d-c. voltages of any magnitude without the consumption of power from the unknown potential. The voltages that can be handled directly by such a device appear to be of almost unlimited magnitude. An inverted vacuum tube made with a small plate electrode at one end of a tubular bulb, with the filament and grid at the other end, would stand very high potentials, and could be made to have almost any desired step-down ratio. The construction would be relatively inexpensive even for voltages in the hundreds of thousands.

It is to be noted that an ordinary type vacuum tube with a μ less than unity (obtained by wide spacing of the grid wires) would have the same general properties as the inverted vacuum tube, but would at the same time be somewhat different in its method of operation. In the usual vacuum tube the element controlling the flow of electrons (grid) is placed between the filament and the electrode receiving the electrons, so the electrons must stream through the controlling electrode. In the inverted vacuum tube the controlling electrode (plate) has no electrons going anywhere near it, being situated outside both the filament and the element receiving the electrons (grid). The advantage of the inverted vacuum tube over an ordinary tube with a μ of less than unity is in the ease with which the step-down ratio may be made large, and in the greater insulation clearance distances for the high-potential electrode. The disadvantage of putting the controlling electrode outside is that the electron flow then goes to a grid, which has limited heat-dissipating capacity. Ordinarily the heat that must be handled is relatively small, however.

Conclusions

By interchanging the functions of the grid and plate of the usual vacuum tube a voltage-reducing power amplifier is obtained. The usual vacuum tube acts as a voltage-increasing power amplifier.

The essential constants of the inverted vacuum tube are the voltage step-down ratio m (approximately equal to μ) and the dynamic grid resistance R_g.

The effect on the grid circuit of applying a voltage in the plate circuit is exactly as though this voltage divided by the step-down ratio m had been added to the grid potential. This equivalent voltage acts in a circuit consisting of the dynamic grid resistance and the output impedance in the grid circuit.

The constants of the inverted vacuum tube can be measured by the simple dynamic means indicated in Figs. 8, 9, and 10.

The step-down ratio m is determined primarily by the geometrical properties of the tube, but at low negative plate potentials and at high grid currents it changes somewhat with operating voltages. At higher negative plate potentials and at moderate grid currents the step-down ratio is practically constant, and is approximately equal to the μ of the tube.

The dynamic grid resistance of the inverted vacuum tube is much lower than the dynamic plate resistance of the same tube used in the usual manner. This low resistance is obtained with low grid potentials, and seems to be largely independent of spacing and size of the grid wires.

The inverted vacuum tube can be used as a voltage amplifier, oscillator, rectifier, or as a voltmeter for the measurement of either a-c. or d-c. voltages of high values with practically no consumption of power. Another use is in oscillograph work, for the taking of voltage oscillograms in high-voltage circuits without consumption of power.

The input resistance of the inverted vacuum tube is practically infinite, while the input capacity is very small, ordinarily not over 5 to 20 $\mu\mu$fds.

The static curves of the inverted vacuum tube are similar in form to the corresponding curves of the ordinary vacuum tube, and the theory of the inverted vacuum tube is analogous in all respects to the usual vacuum-tube theory, the only difference being reduction instead of amplification of voltage.

It is relatively simple to construct an inverted vacuum tube with wide clearances between plate and the rest of the tube, so that potentials of hundreds of thousands of volts can be applied to the plate, while the effect of this high voltage stepped down in almost any desired ratio is obtained in a low-potential circuit.

DEVELOPMENT OF A NEW POWER AMPLIFIER TUBE*

By

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(Westinghouse Research Laboratory, East Pittsburgh, Penna.)

Summary—A general rule for determining the best operating point and load impedance for any power amplifier when anode voltage and dissipation limits must both be considered is derived.

The effect of varying the voltage factor in a given sized tube by changing the grid structure is also considered with voltage and heating kept within safe limits. The process of determining the desired characteristics for a given application is illustrated by describing the development of a new power tube, Radiotron UX-250.

Introduction

In the several papers which have appeared during the last few years on power amplifiers,[1] the output characteristics of existing standard tubes were considered. This paper has for its purpose to describe the process of development of a new power tube known as Radiotron UX−250 for radio receivers, electric phonographs and other applications requiring greater power than obtainable in previous tubes employing standard receiving tube bases. Although the development of this new tube is the principal consideration of the paper, certain properties and limitations as well as design information which apply to power tubes and their associated circuits in general are shown.

Determination of Best Load Impedance and Operating Point

It will not be out of place first of all to review the methods of determining the load impedance and operating point for obtaining the maximum power from a given amplifier. The only case which will be considered is the one in which the anode voltage and dissipation have fixed upper limits, and the grid bias and load impedance may both be chosen so as to obtain maximum undistorted output from the amplifier circuit. This means maximum power without regard to sensitivity, the only consideration being that the curvature of the dynamic characteristic shall be within a certain limit. This curvature produces distortion which is

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* Presented at New York Institute Meeting, March 7, 1928.

[1] E. W. Kellogg, "Design of Non-Distorting Power Amplifiers" *Journal* A. I. E. E., May, 1925. J. C. Warner and A. V. Loughren, "Output Characteristics of Amplifier Tubes," Proceedings I. R. E., December, 1926.

most objectionable on a combination of several musical tones because the various frequencies modulate one another, producing sum and difference frequencies which are inharmonic. A simple measure of the extent of such modulation is the percentage of second harmonic introduced due to curvature when a single fundamental frequency is applied to the grid. An arbitrary value of 5 per cent as an allowable maximum for the second harmonic has been agreed upon by several workers in this field.

In all of the determinations given in this paper, the circuit of Fig. 1 or its equivalent will be considered. In this circuit all of the direct current in the anode circuit is considered to pass through the low-loss reactor L without appreciable voltage drop. The

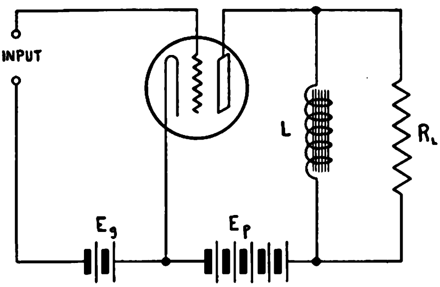

Fig. 1—Power Amplifier Circuit

a-c. output is considered to flow entirely through the load resistance R_L, because the impedance of the reactor to alternating current is very high. A load connected to the tube through a transformer is of course equivalent to the foregoing if the transformer exciting current is very small.[2] If a transformer is used, the equivalent load impedance is the secondary load resistance multiplied by the square of the turn ratio. Because of the comparative ease in obtaining transformers of different ratios, it will be considered that the equivalent load resistance may be made any value desired.

The best load impedance and operating point will now be determined. It was shown theoretically by W. J. Brown[3] that the maximum undistorted output of any tube at a given mean anode

[2] A direct method of designing transformers and reactors capable of carrying considerable direct current and having high excitation inductance is given in a recent paper by C. R. Hanna on "Design of Reactances and Transformers Which Carry Direct Current." *Journal* A. I. E. E., March, 1927.

[3] Discussion, Symposium on Loud Speakers, *Proceedings* of London Physical Society, 36, Part III, April 1, 1924.

voltage will be obtained if the load impedance is equal to twice the tube impedance. In some cases this condition requires that the mean anode current shall be so high as to cause excessive heating of the anode. It will be shown here that under these conditions the maximum permissible anode dissipation should be employed and the load impedance should be greater than twice the tube impedance.

Since the proof given by Mr. Brown that the load impedance should be twice the tube impedance where anode voltage is the only limitation was entirely theoretical, it was thought well to establish this fact experimentally. The curves of Fig. 2 represent the mutual characteristics of a particular tube. Suppose it is

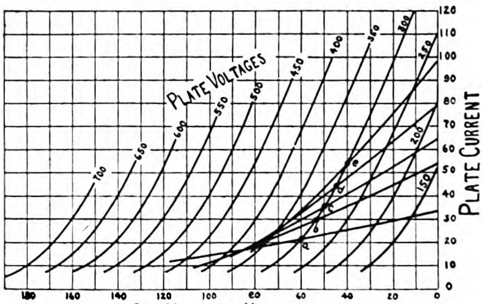

Fig. 2

desired to find the best operating point on the 300-volt curve Through each of the several points, *a, b, c, d,* and *e* on this curve, dynamic characteristics having curvature sufficiently small to prevent the second harmonic from exceeding 5 per cent may be plotted. To do this, a load impedance is arbitrarily chosen and points where the dynamic crosses the various static curves are determined by knowledge of the fact that the change in anode current from one static curve to an adjacent static curve is equal to the difference in anode voltage divided by the load impedance. This fact is seen from the circuit of Fig. 1. When the grid voltage is made more positive the current through the tube must increase. By hypothesis this increase in anode current must all pass through the load resistance, and the drop through this resistance is in such a direction as to lower the anode potential. When the grid is made more negative the anode current is reduced and steady current through the reactor must divide so that part passes through the load resistance in the reverse direction. The voltage drop pro-

duced by this current is in such a direction as to make the anode voltage higher than the supply voltage. In reality the reactor causes the voltage at the anode to rise to this value, since any tendency for the reactor current to be less results in a change of voltage which is great enough to keep the reactor current steady. Thus the dynamic curve intersects static curves corresponding to anode voltages both higher and lower than the supply voltage.

The dynamic curve is, of course, extended to the left to a point where the grid voltage is equal to twice that at the operating

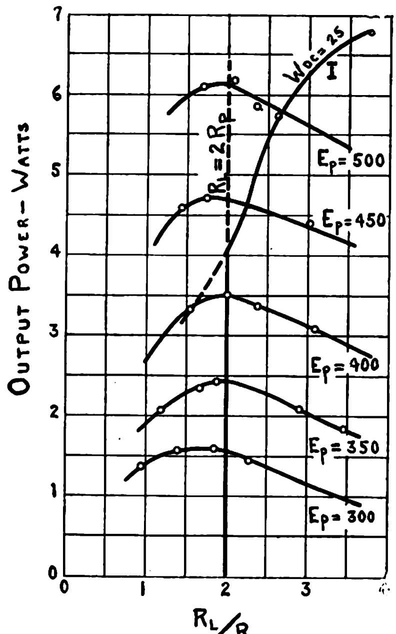

Fig. 3

point and to the right to the point where the grid voltage is zero. Having plotted this dynamic, it remains to be seen whether the curvature is sufficiently small, and this is determined by a calculation of the second harmonic of current using the formula given by Kellogg.[1]

$$\text{Distortion} = \frac{1/2\ (I_{max} + I_{min}) - I_o}{I_{max} - I_{min}}$$

where I_{max} is the anode current for zero grid voltage, I_{min} is the anode current at the lower end of the dynamic and I_o is the anode

current at the operating point. In case this percentage is greater than 5, a larger value of load impedance should be chosen. After having found the dynamic whose curvature is within the limits of allowable distortion, the power may be calculated with the use of the equation

$$W = 1/8\ (I_{max} - I_{min})^2\ R_L$$

The several dynamic curves shown in Fig. 2 are plotted with the correct value of load resistance so as to make the distortion 5 per cent.

In Table I are given the amounts of power obtainable at each of these operating points. The load impedance and the tube impedance, and their ratio at these various operating points, are also tabulated.

TABLE I

E_p	E_g	I_o	R_p	R	W	R_L/R_p
300	59	0.022	3400	17500	1	5.15
"	53	0.030	2670	6050	1.43	2.26
"	49.5	0.036	2340	4400	1.6	1.88
"	45	0.044	2050	2870	1.56	1.40
"	40	0.0545	1800	1700	1.35	0.945

The 300-volt curve of Fig. 3 shows the variation of power plotted against the ratio of load impedance to tube impedance. It is seen to have a maximum where R_L/R_p is approximately 2. Data similar to that given in Table I were also obtained for other anode voltages, and the curves showing power against R_L/R_p for each of these anode voltages are also shown in Fig. 3. Although these maxima are all quite broad, indicating that close matching is not necessary, they occur where R_L/R_p is approximately 2 and therefore substantiate the theoretical conclusions of Mr. Brown.

Curve 1 of Fig. 4 intersects the various static curves for the same tube at the best operating points if anode dissipation is not limited. Dynamic curves are shown through each of these operating points. As the mean anode voltage is increased, however, the mean anode current also increases and somewhere a limit of anode dissipation will be reached. It might be thought that higher voltages would be undesirable, since the mean anode current must be made less and less if a maximum value is set for the anode dissipation.

Curve II of Fig. 4 intersects the anode voltage curves at points where the anode dissipation is constant and, in this case, is arbitrarily fixed at 25 watts. Dynamic curves plotted through these several points for voltages above 425 require load impedances whose values are greater than twice the tube impedance in order

to keep the curvature within the specified limits. It remains to be seen whether the power increases for higher voltages in this region where the mean anode current must necessarily be made smaller and smaller in view of the constant anode dissipation. Curve I of Fig. 3 is for 25 watts anode dissipation and shows that the power

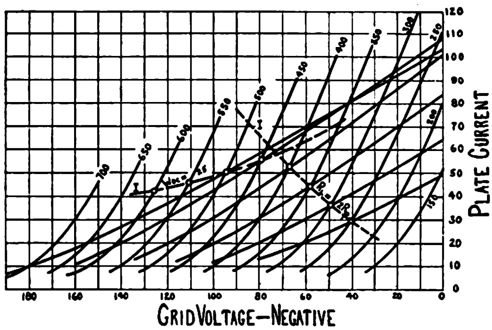

Fig. 4

does steadily increase even under these circumstances, and also shows that R_L/R_p increases with increased anode voltage.

It is seen from the foregoing that in some cases it is desirable to operate at the point which requires that the load impedance shall be twice the tube impedance, and in other cases that the

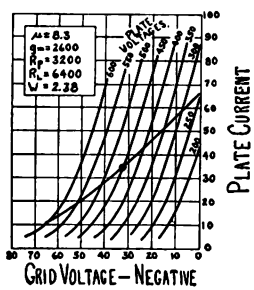

Fig. 5

load impedance shall be greater than twice the tube impedance. A general rule for determining the best operating point where both anode voltage and anode dissipation are limited may be given as follows: On the curve for the greatest permissible anode voltage choose the point where the current is such as to cause the maximum allowable anode dissipation. Plot a dynamic curve by the method previously described, using a value of load resistance great enough to keep the distortion small. If this load impedance is greater than

twice the tube impedance at the point the best operating condition will have been determined. In case the load impedance so chosen is less than twice the tube impedance, greater undistorted output will be obtained if the plate current is reduced by the use of greater grid bias, and a larger value of load impedance employed,

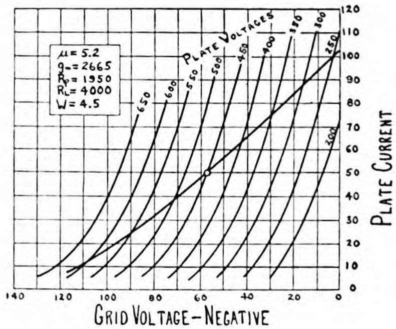

Fig. 6

the best condition resulting when the load impedance is twice the tube impedance.

Variation of Power With Voltage Factor

Having established a simple general rule for determining the best operating point and power output of a given tube where

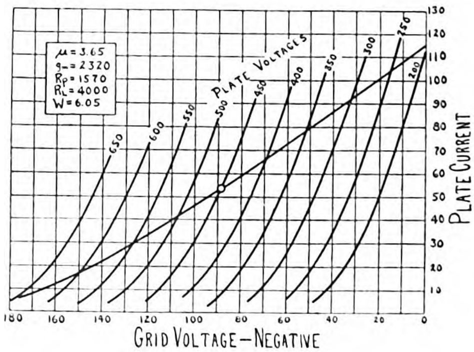

Fig. 7

anode voltage and anode dissipation limits are both taken into consideration, an investigation of the variation in power with voltage factor in a tube of a certain class will be determined. The particular tube under consideration is one with a standard receiving tube base having as large a glass blank as feasible for this size

of base and operating at as great an anode voltage and anode dissipation as would be tolerable in a structure of this type. The voltage limitations are determined largely by press and base structure, while the dissipation is determined by the glass envelope and the size of parts that can be enclosed. It was felt that a tube of this class could be made which would be capable of operating at 450 volts mean anode potential and 25 watts dissipation. A number of tubes having a particular anode and filament structure, but different grid pitches, were made up and the best operating points for the 450-volt condition were determined.

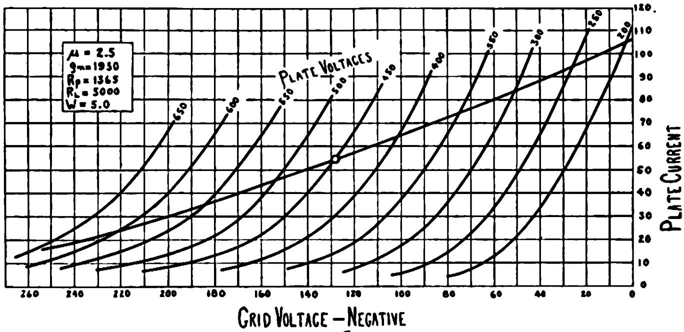

Fig. 8

The curves of Figs. 5, 6, 7, and 8 show the static and dynamic characteristics for four of the tubes, the voltage factors being 8.3, 5.2, 3.65 and 2.5.

A tabulation of data obtained from these and other representative tubes having voltage factors in this range is given in Table II. Several things are apparent from a study of this table. The higher voltage factor tubes operate best with the dissipation

TABLE II

Amplification Factor	Mutual Cond.	Plate Resist. (Ohms) R_p	Load Resistance R_L	D-C Plate Current (Milliamperes I_p	Anode Dissipation in Watts W_{d-c}	Undistorted Output Watts W_p	Negative Grid-Bias (Volts) E_g	Sensitivity $\frac{M.W.}{E_g^2}$
8.3	2600	3200	6400	35	15.8	2.38	32	2.32
7.4	2780	2660	5320	40	18.0	3.08	38	2.13
6.34	2700	2350	4700	45	20.3	3.62	46	1.71
5.2	2665	1950	4000	50	22.5	4.5	58	1.33
3.65	2320	1570	4000	55	25	6.05	88.5	0.77
3.1	2080	1490	4500	55	25	6.2	106	0.55
2.5	1830	1365	5000	55	25	5.0	127.5	0.31

well under 25 watts and load impedance equal to twice the tube impedance. As the voltage factor is decreased, the anode dissipation becomes greater and the tubes having voltage factors of 3.65

and below require that the anode current be kept constant at 55 milliamperes in order not to exceed the limit of 25 watts which has been set. It is also seen from the table that the power output is greater for the lower voltage factor tubes, even below the one where the anode current becomes great enough to cause the anode dissipation to reach its limit. For the very low voltage factors, however, the output is considerably less, as will be noted for the tube having a voltage factor of 2.5. This is due to the poor control of the grid at low anode currents, which may be observed from the static curves of Fig. 8. Thus it may be expected that a limitation will always be reached when the grid structure is made more open. The lower limit of practical voltage factor will, of course, be different for different types of tubes.

Another fact which should be pointed out is that for the lower voltage factor tubes, in the region where constant anode dissipation obtains, the best load impedance steadily increases even though the tube impedance decreases.

It is, of course, to be expected that the sensitivity will be less for the lower voltage factor tubes, and in the case under consideration a limit was reached because of the fact that the maximum grid swing available for operating the tube was about 80 volts. This was the value estimated for the undistorted voltage of a detector and one stage of amplification following.

Curve I of Fig. 9 shows the variation of power with voltage factor. Curve II of Fig. 9 shows the peak grid swing required to give maximum output. The vertical line corresponding to the voltage factor of a tube which requires 80 volts grid swing is drawn. The value of μ is 3.8. It is seen that further decrease in voltage factor with corresponding increase in bias causes only small gain in power output. The dotted curve in Fig. 9 shows what the output would be if there were no anode dissipation limit. In this particular case the maximum power is not greatly in excess of that obtainable when anode dissipation is limited, and the drooping in the low voltage factor region still occurs. This experimental fact of an actual loss in power when the voltage factor is decreased below a certain value in a given size tube was not brought out in the curves shown by Messrs. Warner and Loughren in Fig. 10 of their paper.[1] Their curves must have been derived on theoretical grounds with the assumption that the mutual conductance for very low anode currents did not decrease excessively with decrease in voltage factor. In practice,

tubes of very low voltage factor have such poor grid control at the lower end of the dynamic curve as to necessitate a higher value of I_{min} in order to avoid excessive curvature. The resulting smaller anode current swing causes a decrease in power output even though the required load resistance is larger.

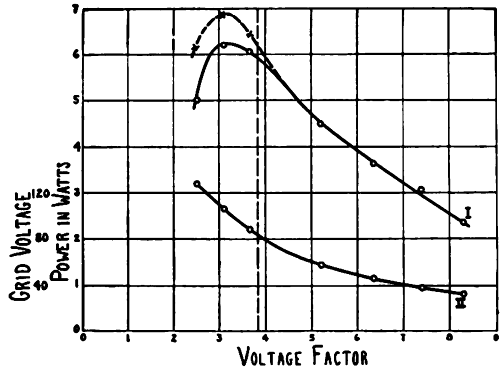

Fig. 9

In view of the 80-volt limitation set for grid swing, the tube finally chosen for the application in mind was one having a voltage factor of 3.8. The average of tubes made in the factory have a

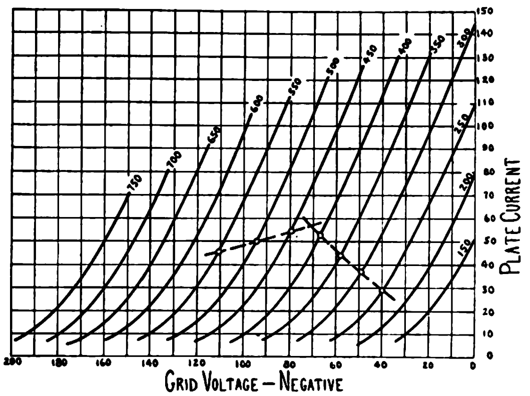

Fig. 10

mutual conductance of 2100 micro-ohms and an anode impedance of 1800 ohms. Fig. 10 shows the mutual characteristics of this tube. Correct operating points at various plate voltages are indicated. Fig. 11 shows power output available at different plate voltages. It will be seen that the power at the normal operating

point of 450 volts plate and 80 volts grid is 4.6 watts. This represents an average tube as made in the factory and is known as the UX-250.

Description of the UX-250

The UX-250 tube is shown in Fig. 12. The tube has the large X base, which is standard in receiving sets. A 2–5/8 in. diameter

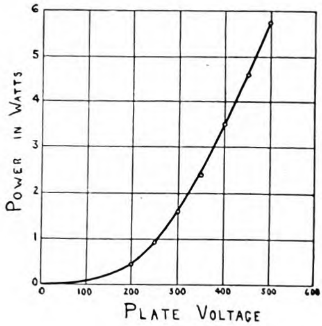

Fig. 11

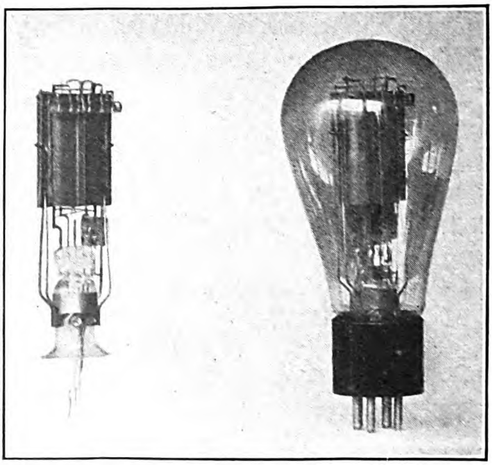

Fig. 12—The UX-250—A New Power Receiving Tube

glass blank is employed, giving an overall height for the tube of 6–1/4 in. The anode is of oval construction with 1/4 in. cooling fins on each of the two flat faces, and in addition has a black surface to facilitate cooling. Its dimensions are 5/16 in.$\times$13/16 in. $\times$1–5/8 in. long. The filament is of the oxide-coated type of very

high efficiency, requiring only 1.25 amperes at its rated voltage of 7.5. The grid as well as the plate of this tube is black because of a special treatment which makes possible its functioning also as an oscillator.

A tabulation of power output characteristics and operating points for various Radiotron power tubes employing receiving ube bases is given in Table III. The data for the UX-112A and

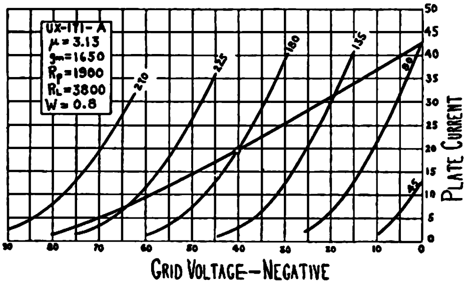

Fig. 13

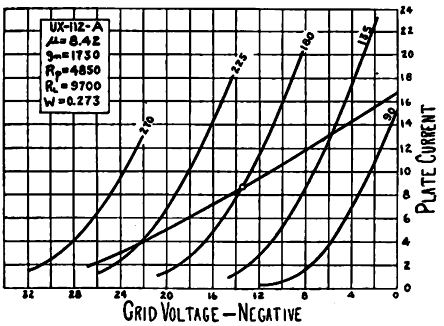

Fig. 14

UX-171A were obtained from the curves of Figs. 13 and 14. Data for other tubes except the UX-250 were taken from the paper by Warner and Loughren.[1]

TABLE III

Tube	Plate Volts E_p	Grid Volts E_g	D-C Plate Current (Milli-amperes) I_o	Plate Dissi-pation Watts $W_{d\text{-}c}$	Voltage Factor	Plate Resis-tance R_p	Load Resis-tance R_L	Max. Output (Watts) W
UX-250	450	80	55	25	3.8	1800	4000	4.6
	400	67.5	52.5	21	3.8	1850	3700	3.5
	350	58.5	44	16.5	3.8	2050	4100	2.45
UX-210	400	35	16	6.4	7.5	5400	11000	1.34
UX-171A	180	40.5	19	3.42	3.1	1900	3800	0.8
UX-112A	180	13.5	8.5	1.54	8.4	4850	9700	0.273
UX-120	135	22.5	7	0.942	3.3	6600	13200	0.105

The high output of the UX-250 is the result of several features, the most important of which are enumerated below:

(1) The use of as large a glass blank as feasible on a receiving tube base so as to allow high plate dissipation.

(2) The use of a large plate with a black surface for good radiation at a safe temperature.

(3) The use of the oxide-coated filament for ample emission without excessive filament power.

(4) The choice of a fairly open grid structure to obtain low impedance.

It is probable that the output of this tube is about as great as may be obtained from a receiving tube of practical structure without requiring greater grid swing or operating at excessive anode voltage or dissipation. The UX-250 is useful in the many applications where very high voltages cannot be used, but large amounts of power are required.

Discussion*

J. C. Warner and A. V. Loughren:† This paper raises the interesting question of the decrease in amplification factor at low plate currents in a given tube. This is particularly noticeable in a tube designed to have a low amplification factor when the grid wires are relatively far apart. If the spacing between the grid wires becomes too large compared with the distance from the grid wires to the cathode, the grid control obviously cannot be uniform at the cathode. This results in a decreased amplification factor when the grid is made more and more negative, since the current coming from parts of the cathode directly opposite the grid wires tends to be cut off first.

In the experiments described in this paper it is stated that the amplification factor was reduced by increasing the distance between grid wires. In this case the distance between the wires becomes equal to the distance from grid to cathode and anode at an amplification factor of about 3.5. Any further increase in spacing is accompanied by serious loss of control. However, if the reduction in amplification factor is accomplished by a reduction in grid-wire diameter instead of spacing, an amplification factor of less than 2 may be reached without making the wire spacing greater than the distance to the cathode.

*Original Manuscript Received by the Institute, March 19, 1928.
†General Electric Company, Schenectady, N. Y.

Of course for practical reasons it is not always possible to use the small grid wires required in the second illustration, but this example is given merely to indicate the possibility of designing a tube, when occasion demands, to have a practically constant value of amplification factor even when this value is relatively low.

In connection with the change in mutual conductance with varying amplification factor (varied by changing design of grid mesh) it should be emphasized that when the plate current is held constant the mutual conductance always decreases with the amplification factor. This shows very clearly that in comparing different tubes the mutual conductance tells nothing whatever regarding the maximum undistorted output which may be obtained.

There is an error in the data given for the UX-120 in Table III. The load resistance should be 6,600 ohms instead of 13,200. This is one of the cases where for practical reasons the tube design and operating conditions are worked out to give maximum output into a load resistance equal to the plate resistance instead of twice as great. The theoretical optimum conditions would require a load resistance rather hard to obtain in practice and the output would be only about 5 per cent greater than before.

MEASUREMENT OF VACUUM-TUBE CAPACITIES BY A TRANSFORMER BALANCE*

By

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Summary—*A complete, portable equipment is described for the measurement of the direct capacities of vacuum tubes in laboratory or factory testing. The tube capacity is compared with a standard variable condenser by means of a transformer-balance (Neutrodyne) circuit, whose balance is independent of the frequency (about 1500 kc. being preferred). Designs are proposed for the standard condenser and the transformer, and suggestions are made for the further improvement of this equipment.*

Much attention is being directed at present to the precise measurement of the small inter-electrode capacities of vacuum tubes, especially the grid-plate capacity, which is on the order of ten micro-microfarads. A method of measuring this capacity is required which will give reproducible results within about one percent when assembled in any laboratory.

This problem has arisen with the commercial development of Neutrodyne receivers, dating from 1923. Prior to this time, amateur and broadcast receivers as a class were designed and built with relatively wide manufacturing tolerances, and vacuum tubes evolved in this class. The Neutrodyne receiver involves in each radio-frequency amplifier tube a *balance* between the grid-plate capacity and the capacity of an added condenser, which is adjusted in the factory. As the earlier sets and vacuum tubes were improved, it was soon discovered that the production variations in grid-plate capacity were large enough to require neutralization for individual tubes. Since this procedure was ruled out by expediency the only alternative was to reduce the amplification to the point where the set could be neutralized for average tubes and subsequently operated with any tubes of a given design. Since this became a serious limitation, the problem was studied from several angles. First, the designing of sets required a knowledge of the range of grid-plate capacities to be met with in tubes on the market. Secondly, the neutralizing of manufactured sets required the selection and use of tubes with average grid-plate capacities. And thirdly, the manufacturing of tubes required the accurate specifications and checking of the grid-plate capacity in order to reject tubes with deviations greater than the specified tolerance.

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* Presented at the Annual Institute Convention, January 11, 1928.

Since a tube has three direct inter-electrode capacities (grid-plate, grid-filament, and plate-filament) in a delta connection, it is not easy to measure one of these by a single observation. Three observations would be required unless an arrangement were used in which the effects of the other two capacities could be neglected. Various methods have been proposed to comply with this condition.

It is the purpose of this paper to outline an equipment for measuring the inter-electrode capacities of a tube by means of a tuned radio-frequency transformer balance patterned after the

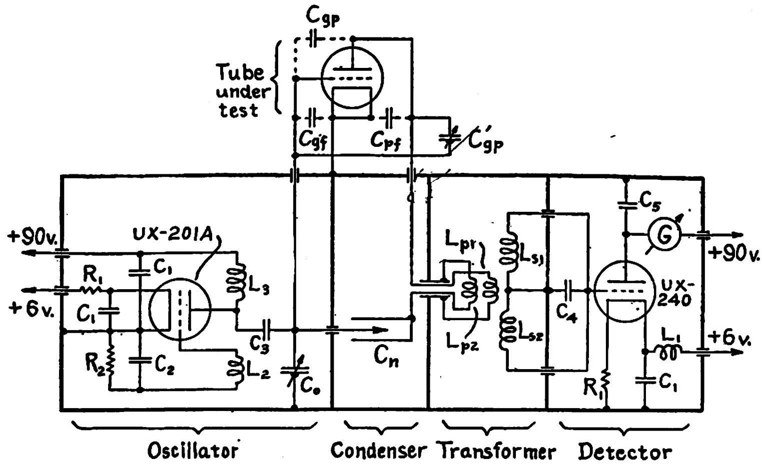

Fig. 1.—Complete Transformer-Balance (Neutrodyne) Circuit

Neutrodyne circuit. This is a logical solution of this problem which arose in the development of Neutrodyne sets, and will be seen to have some advantages over other proposed schemes.

The complete circuit arrangement is shown in Fig. 1. It is housed in a copper box divided into four compartments, and comprises five component parts,—a radio-frequency oscillator, the tube under test with its socket, a standard variable condenser, the radio-frequency transformer, and a detector circuit. These will be described individually in order. The balance circuit comprises the two capacities, C'_{gp} and C_n, and the two like transformer primaries, L_{p1} and L_{p2}, which are arranged as in the Neutrodyne circuit. When $C_n = C_{gp}$ the net voltage induced in the transformer secondary is zero. Measurements are made at a frequency of about 1500 kc. The battery circuits are filtered to permit the use of common batteries for oscillator and detector.

The oscillator, located in the first compartment, employs a simple circuit with a UX-201A tube. The power in the tuned circuit is about 50 milliwatts at about 50 volts. The variable condenser C_0 of about 200 $\mu\mu$f. maximum capacity serves to tune the oscillator to resonance with the detector circuit at approximately 1500 kc. The plate coil L_3 consists of 50 turns of No. 24 single-silk enamel wire on a form 3 in. in diameter, while the grid coil has 20 turns on a smaller form just under the ground end of the plate coil. $C_1=1$ μf., $C_2=0.001$ μf., $C_3=0.01$ μf., $R_1=4$ ohms, $R_2=0.1$ megohm.

The tube under test is mounted in a special socket, such as may be agreed upon, conveniently located on top of the box. The incidental capacities between electrodes are kept as small as pos-

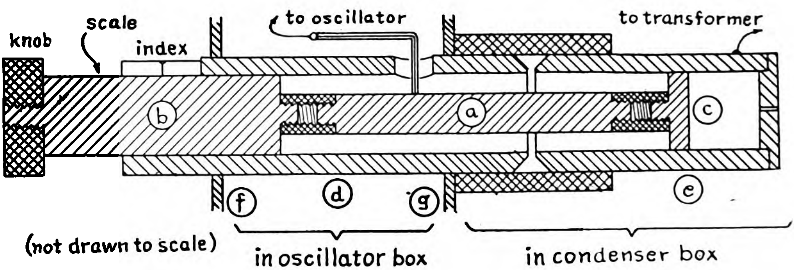

Fig. 2.—Standard Variable Condenser

sible. A small adjustable capacity C_{gp}' may be provided as a zero adjustment when the socket is readjusted for different types of tubes.

The variable condenser C_n, in the second compartment, is the standard of comparison for the measurements. It occupies the position of the neutralizing condenser in the Neutrodyne circuit. This condenser can best be composed of two coaxial cylinders, one sliding to vary the capacity. A proposed design is shown in Fig. 2. A brass rod *a* is supported by insulating sleeves between two pistons *b* and *c*. These pistons slide in two sections of cylindrical tubing, *d* and *e* respectively, which are fastened end to end by an insulating sleeve. The tube *d* extends from outside the box through the wall *f* into the oscillator compartment and then through the partition *g* into the condenser compartment. The piston *b* carries a knob and a scale which passes an index on the outer tube *d*. Electrical connection from the oscillator to the sliding rod *a* is made by contact between the rod and an insulated spring brush entering through

tube d, which is grounded to the box. The standard variable capacity is between the rod a and the section of outer tube e, which is connected to the transformer. The capacity between e and the box should be kept small.

This condenser design has several desirable mechanical and electrical features. The outside diameter of the rod a and the inside diameter of tubes d and e can be turned to such a ratio (0.573, or approximately 4/7) that one centimeter on the scale represents one micro-microfarad of capacity. The entire assembly is a unit in the outer tubes, and need not be calibrated experimentally except as a check on computations, or for higher accuracy. The computed calibration is realized very closely by avoiding two

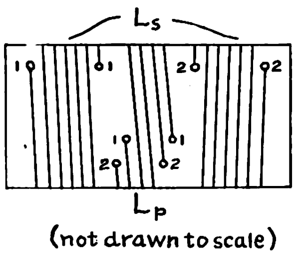

Fig. 3.—Transformer Arrangement

end corrections of the capacity between a and e. First, a radial electric field at the open end of e is secured by the use of d as a "guard cylinder." Secondly, the variable capacity between a and the closed end of e is avoided by closing e on the inside with c, which moves with a. The only remaining error is the distortion of the radial field at the end of a; so that a must remain within e by a distance greater than the diameter of a. By this design, all variations of capacity are completely shielded from effects of outside objects. Another advantage of this design is that the total capacity from a to ground directly and through the transformer, is constant. After the oscillator is tuned to the detector, with the tube in the socket, the condition of resonance is not altered by sliding a to vary C_n. Retuning for different tubes of the same type should be unnecessary, so that the oscillator requires tuning only once.

The transformer is enclosed in the third compartment, and the arrangement of its coils on a single form is shown in Fig. 3. The two primary coils, L_{p1} and L_{p2}, each consist of 10 turns of No. 24 single-silk enamel wire, and their turns are interleaved. The two

secondary coils, L_{s1} and L_{s2}, each consists of 50 turns of No. 24 single-silk enamel wire, and they are wound on opposite sides of the primary coils in opposite directions as shown. The winding tube is of clear bakelite, 3 in. diameter by 5 in. long, and should be threaded 18 double threads per inch for primary and 36 threads per inch for secondary coils. The symmetry is maintained as precisely as possible in the coil construction, and the assembly with its connecting wires is located symmetrically in the compartment, as indicated in Fig. 1. The primary leads are brought to the coils in a grounded metal tube, which serves as the return to ground for the primary coils only. The inner secondary terminals are grounded, while the outer terminals are connected in parallel in the detector compartment. This arrangement minimizes all incidental coupling, capacitive or inductive, between primary and secondary coils.

The detector is located in the fourth compartment, as shown in Fig. 1. A UX-240 tube is used, with a grid bias of one volt secured by the filament resistance R_1 of 4 ohms. The grid is connected to the transformer secondaries, which are tuned to about 1500 kc. by a small condenser C_4 of 50 $\mu\mu$f., augmented by incidental capacities. This circuit is tuned to increase the sensitivity only. High selectivity is not desirable here. The galvanometer G has a full scale deflection for about 0.5 milliampere, and indicates relative voltages on the grid. The battery circuits are filtered by $C_5=0.1$ μf., $C_1=1$ μf., $L_1=0.1$ mh.

The total errors of this system as described should be very small, within 1 percent or 0.1 $\mu\mu$f., and can be reduced further if required. The mechanical errors in the standard condenser and transformer can be reduced almost indefinitely. The principal source of electrical errors is in the transformer primary coils, because it is impossible to approach very closely to unity coupling between two coils of so few turns. Any deviation from symmetry between these two coils can be checked by reversing the primary leads in the second compartment. Other errors in the circuit vary with the frequency and can be detected by comparing measurements at different frequencies with the same transformer. These errors can be reduced by decreasing the number of primary turns or working at a lower frequency.

Taken as a whole, this system has several definite advantages which are fundamentally important. The null method of measurement is favored because it obviates the accurate calibration and

checking of meters, and because variations in the applied voltage do not detract from the accuracy of the measurements. In this case, even the comparison standard should not require calibration if carefully computed and constructed. Then this transformer balance is independent of the frequency, so that constancy of the oscillator frequency is not essential. This tolerance of variations in oscillator voltage and frequency permits the use of a low-power oscillator, with the result that the equipment is light and portable. Also the equipment can be operated without unnecessary delays, because the tubes need not come to equilibrium after being lighted and because there is no time lag in the galvanometer circuit as contrasted with thermal instruments. The use of a radio frequency is to be preferred over an audio frequency, since for the latter the impedance of such small capacities is very high, and also the difficulties of oscillator design are increased.

In high precision laboratory work, this system offers indefinite opportunities for further refinement by improving the transformer design and construction, and by using a radio-frequency amplifier before the detector. It is especially adaptable to the measurement of the minute coupling capacities encountered in screen-electrode tubes. The variable condenser can be improved by the use of a micrometer screw. The highest accuracy is attainable by taking the mean of two settings equally spaced on opposite sides of balance.

In factory testing, this system has the advantage that an error in capacity either above or below normal gives a rapid increase in galvanometer current, and a relay can be operated to show when tubes should be rejected for such errors. The absence of time lag in the detector is especially important in this class of work.

A DIRECT-CAPACITY BRIDGE FOR VACUUM-TUBE MEASUREMENTS*

By

Lincoln Walsh

(Consulting Engineer, Elizabeth, N. J.)

Summary—A direct-capacity bridge is described which permits the measurement at a single setting of a capacity associated with other capacities in a system having more than two terminals, such as the grid-plate capacity of a vacuum tube. Two forms of the bridge are described. By making one connection the standard form of capacity bridge already in use in many laboratories may be converted into a direct-capacity bridge. The recommendation is made that vacuum-tube inter-element capacities be specified as direct capacities. Suggestions are made for other uses of the direct-capacity bridge in the laboratory.

THE object of this paper is to present a modification of the standard capacity bridge which permits the measurement of direct capacities, and which has been found suitable for measurement of vacuum tubes.

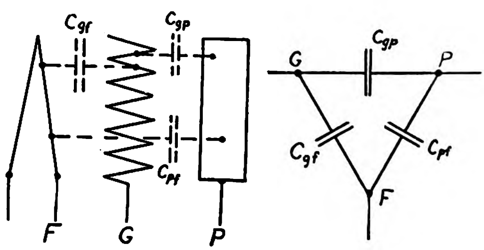

Fig. 1

The direct capacity between two elements of a system such as the vacuum tube may be described as the capacity existing between those two elements, excluding all other capacities of the system.[1]

In the vacuum tube there exist three direct capacities, grid-plate, grid-filament, and plate-filament, the system having three terminals, grid, plate, and filament, as shown in Fig. 1. The standard capacity bridge is adapted to measure the capacity of two-terminal systems only, such as the single capacity between the two terminals of a condenser, and cannot measure directly the individual direct capacities of the vacuum tube.

To meet the need for apparatus capable of measuring the direct capacities of vacuum tubes, the bridge to be described was developed in the Hazeltine Laboratories by the writer.

* Original Manuscript Received by the Institute, December 14, 1927.

* Presented at the Annual Institute Convention, January 11, 1928.

[1] See definition 4031 of the "Report of the Standardization Committee for 1926"

The circuit of the direct capacity bridge is shown in Fig. 2. The capacity to be measured, C_{gp}, is connected between corners A and D of the bridge and the terminal or terminals not associated with the capacity to be measured are connected to the junction of the ratio arms, which is grounded. C_s is the standard of capacity, and r_m the phase-angle resistance. The capacity C_{pf} is thus placed across the bridge from B to D, where it is effectively

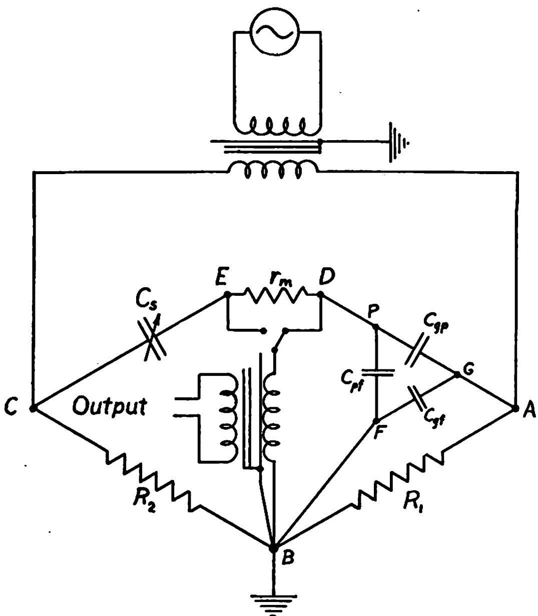

Fig. 2

outside the bridge and does not in any way affect the balance. The capacity C_{gf} is in parallel with the resistance arm R_1. Fig 3 shows the bridge in simplified form.

When the bridge is balanced,

$$\frac{Z_2}{Z_1}=\frac{Z_4}{Z_3} \text{ or } \frac{R_2}{\dfrac{1}{\dfrac{1}{R_1}+j\omega C_1}}=\frac{R_4+\dfrac{1}{j\omega C_4}}{\dfrac{1}{j\omega C_3}}.$$

This reduces directly to

$$\frac{R_2}{R_1}(1+j\omega C_1R_1)=\frac{C_3}{C_4}(1+j\omega C_4R_4),$$

which leads to the two requirements, that

$$C_1R_1 = C_4R_4$$

and that

$$\frac{R_2}{R_1} = \frac{C_3}{C_4}.$$

The result shows that the only effect of the capacity introduced in parallel with R_1 is to require resistance to be inserted in series with the standard capacity to balance it, in addition to the resistance needed to balance the loss in the capacity being measured, and that it does not affect the ratio of the unknown to the standard

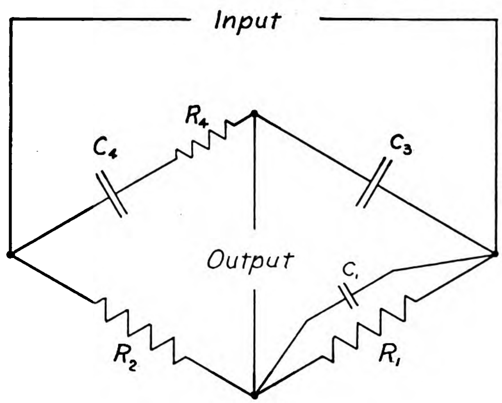

Fig. 3

capacity. Therefore the bridge as described may be used to measure direct capacity.

Only one setting is needed to measure a direct capacity after the "zero" setting of the standard condenser has been found. The procedure is the same as that followed on the standard bridge.

The only difference between the direct capacity bridge and the standard bridge is the connection between the grounded junction of the ratio arms and the terminal or terminals not associated with the capacity being measured. Therefore any standard bridge can be readily converted into a direct capacity bridge by providing a grounded terminal to which the non-associated terminal or terminals of the capacity system may be connected.

The rule to be followed in connecting apparatus for direct capacity measurements is:

Connect to the unknown terminals of the bridge, *A* and *D*, the two terminals associated with the direct capacity to be measured.

Connect to ground all other terminals of the capacity system and shields. (It is assumed that the "indicator" terminal B is grounded).

The accuracy of the bridge is limited only by the accuracy of the standard, and may be increased for small capacities by using ratios of 10:1 or 100:1, instead of the usual 1:1. When using these ratios it is generally necessary to use a high-voltage input to the

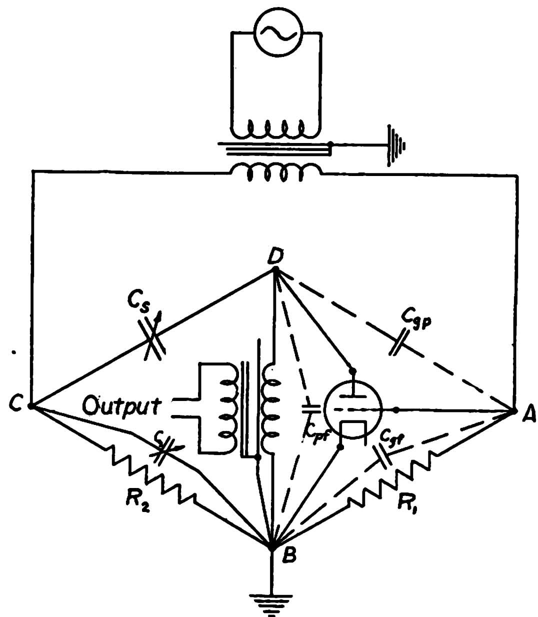

Fig. 4

bridge, and an amplifier on the output, to obtain the sharpest useful balance.

A modification which has appeared in an unpublished report is shown in Fig. 4. The grid-filament capacity in parallel with one of the ratio arms is balanced by a capacity in parallel with the other ratio arm, instead of being balanced by the resistance in series with the standard capacity, as above described. This capacity also serves to balance the bridge for phase angle due to loss in the capacity being measured.

In the wire communication field, the need for accurate measurements of the direct capacities of a multi-capacity system has long been recognized, and apparatus capable of such measurements is in constant use, but this apparatus has not found general application in the radio field. Those who wish to study the subject

further will find this apparatus described in a comprehensive paper on "Measurement of Direct Capacities," by Dr. G. A. Campbell, in the Bell System *Technical Journal* of July, 1922.

Error due to extraneous capacities can be very easily eliminated in direct capacity measurements, so that there is no need to specify exact conditions of mounting in order to secure agreement between various laboratories.

In view of the simplicity and accuracy of apparatus now made available for direct capacity measurements, it is recommended that hereafter all capacities of vacuum tubes shall be given as direct capacities. This is to eliminate the inaccuracy and confusion resulting from the use of total capacities, and capacities measured with one element floating in potential.

Among the uses of the direct capacity bridge in a laboratory, other than measuring vacuum tubes, may be mentioned the measurement of stray coupling capacities; simplification of measurements on dielectrics; and the elimination of error in two-terminal capacity measurements due to stray capacities to leads. by permitting the shielding of the leads.

Discussion

Mr. Walsh: The question has been asked whether or not the use of a Wagner ground would be of advantage on the direct-capacity bridge. The object of the Wagner ground is to eliminate the error caused by capacities from the supply lines to ground, which capacities are effectively in parallel with one or the other of the ratio arms. It was shown in the analysis of Fig. 3 that capacity in parallel with one of the ratio arms does not affect the capacity measurement. In general then, the use of a Wagner ground is not advantageous for any capacity bridge measurements, either total or direct-capacity, except for measurements involving the determination of power factor, in which case its use is always desirable.

The question has been asked as to which of the two forms of the bridge, shown in Figs. 2 and 4 respectively, is preferred. The two forms are equally accurate and satisfactory. The General Radio standard capacity bridge may be converted into the bridge shown in Fig. 2 by providing a grounded terminal, as described in the paper, and as this bridge or something equivalent to it seems to be in general use, the form shown in Fig. 2 was described first.

A BRIDGE METHOD FOR THE MEASUREMENT OF INTER-ELECTRODE ADMITTANCE IN VACUUM TUBES*

By

E. T. Hoch

(Bell Telephone Laboratories, New York City)

Summary—A description is given of the Colpitts-Campbell bridge as applied specifically to the measurement of direct admittances in vacuum tubes. Data are given on several tubes.

IN 1904 G. A. Campbell described a bridge circuit[1] which has proved very useful for the separation of complex electrical networks into their component direct impedances or admittances. In 1922, Campbell discussed further the definition and measurement of direct capacitances.[2] The object of this paper is to describe in greater detail the application of this method to the measurement of the direct capacitances or admittances in vacuum tubes.

As is well known, in a network of three or more terminals the capacitance measured between any two terminals with the other terminal floating, includes not only the direct capacitance between those terminals but also a capacitance consisting of a series parallel arrangement of the direct capacitances to all of the other terminals of the network. For example, the admittances between elements in a three-element vacuum tube constitute a network of six individual admittances as shown in Fig. 1. In this figure, the admittances are shown as capacitances and resistances in parallel although mathematically they may be considered equally well as capacitances and resistances in series. It is the segregation of the individual capacitances shown in the diagram which constitutes the problem of measuring the direct capacitance as compared with the total capacitance between any two terminals.

In Fig. 1, the outer circle marked S represents in general any conductor in proximity to the tube which is not definitely connected to one of the tube elements. In particular, it may represent the earth or, in certain types of tubes, the metal shell surrounding the base. So far as the operation of the tube in a circuit

* Original Manuscript Received by the Institute, January 6, 1928.

* Presented at the Annual Convention, January 11, 1928.

[1] G. A. Campbell, "The Shielded Balance," *El. W.* 43, 1904, (647–649).

[2] G. A. Campbell, "Direct Capacity Measurements," Bell System *Tech. Journal*, 1, 1922 (18–38).

is concerned, the resultant effective capacitance from plate to grid, plate to filament and grid to filament, including the effect of capacitances from each element to S can be determined without attempting to segregate the capacitance from each element to S, providing only that S is connected (or disconnected) the same way during the measurement as it will be when the tube is in use. For design purposes, however, it is sometimes desirable to determine the capacitance to shell or other metal parts independently as shown in the diagram.

It is understood, of course, that the grid-to-filament admittance referred to above is entirely different from the input admit-

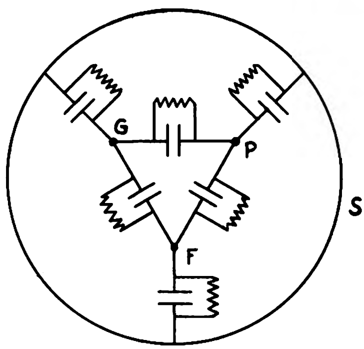

Fig. 1—Network of Admittances in a Vacuum Tube.

tance of the tube under operating conditions since the latter is a function of the plate-circuit impedance and the amplification factor of the tube.[3] The same bridge can, however, be used for the measurement of input admittance under operating conditions.

Description of Bridge

The bridge as originally described by Campbell and as further described by Shackelton and Ferguson,[4] is quite general in its application. For a specific use such as the measurement of small direct capacitances, the details of the bridge may be very much simplified; hence we shall consider only the simplified form shown in Fig. 2.

[3] H. W. Nichols, *Phys. Rev.*, **13**, 1919 (404–419). Miller, Bureau of Standards Scientific Paper No. 351.

[4] Shackleton and Ferguson, "High-Frequency Measurement of Communication Apparatus." Presented at Regional Meeting of the A.I.E.E., Pittsfield, Mass., May 25–28, 1927. See Bell System *Tech. Journal*, Jan. 1928.

As shown in this figure, the bridge proper consists of two shielded transformers one of which should be double shielded, two double shielded equal ratio arms, a differential air capacitor of suitable range, one fixed and one variable resistor. The fixed resistor *CD* is usually of 10,000 ohms resistance. For general purpose work, the *AD* resistor usually consists of six decades having a total resistance of 11,000 ohms variable in steps of 0.01 ohm. For vacuum-tube testing this may consist of a fixed resistor in series with a small variable resistor to give a variation of a few ohms on either side of 10,000 in steps of 0.01 ohm.

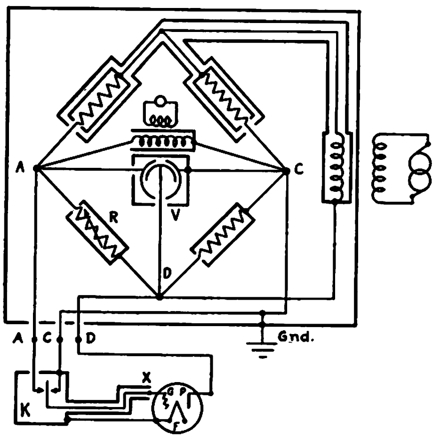

Fig. 2—Bridge Circuit for Measurement of Direct Capacitance and Conductance.

Fig. 2 also shows the connections from the bridge for measuring the direct capacitance from grid to plate of a vacuum tube. The terminals between which the capacitance is to be measured are connected to wires *D* and *X* and the remaining terminal or terminals are connected to *C*. Fig. 2 represents the case of a tube with a metal shell around the base and the shell considered as a separate terminal. If the capacitance with the shell floating is desired, the connection to the shell is omitted. *K* is a shielded key by means of which the capacitance under test may be thrown from the *CD* to the *AD* arm of the bridge or vice versa without disturbing any of the other capacitances in the network.

Procedure

The procedure for making a measurement is as follows: The apparatus is connected as shown in Fig. 2. *K* is set to connect the

capacitance from grid to plate in, say, the CD arm. The bridge is then balanced by adjusting R and V, the readings being R_c and V_c. K is then thrown to the A side and the bridge balanced again, the readings being R_a and V_a. If capacitor V is calibrated in terms of the capacitance which it will balance in the CD arm of the bridge, the direct capacitance is

$$C_{gp}=\frac{V_c-V_a}{2}.$$

The corresponding conductance is

$$G_{gp}=\frac{R_a-R_c}{2R_aR_c}.$$

This may be shown as follows. For the first measurement C_{gp} and C_{pf} are both in the CD arm of the bridge. C_{gf} is short-circuited and so does not affect the bridge balance. Since the bridge is assumed to be calibrated so that capacitances in CD give increasing readings on V, the capacitance in this arm may be considered positive and that in AD as negative. Therefore, for the first measurement the capacitances being balanced are

$$C_{gp}+C_{pf}=V_c.$$

For the second measurement C_{gp} is in AD, C_{pf} is in CD and C_{gf} is across AC and constitutes a shunt across the receiver but has no effect on the balance point of the bridge; therefore the capacitances being balanced on the second measurement are

$$-C_{gp}+C_{pf}=V_a.$$

Solving, we get

$$C_{gp}=\frac{V_c-V_a}{2}.$$

The conductance equation is derived in the same way taking account of the fact that the conductance being measured is added in parallel with the conductance arms of the bridge. If R is normally 10,000 ohms and R_a and R_c do not depart from this value by more than say 100 ohms, G reduces to $\dfrac{R_a-R_c}{200}$ micromhos approximately.

It should be noted that measurements made as described above do not include the capacitances from the tube elements to ground. If it is desired to measure the capacitance of any element to ground independently, corner D of the bridge is connected to ground instead of C as above and the terminal is switched from C to A as before. Likewise, the direct capacitance from one terminal to another terminal which is grounded is made by connecting the grounded terminal to D and switching the other terminal from C to A, all other terminals being connected to C. Therefore, if a vacuum tube is operated with the filament circuit grounded, the capacitances operative in the circuit are obtained by measuring C_{gp} with C of the bridge grounded and measuring C_{pf} and C_{gf} with D grounded and the filament connected to D.

Precautions

The principal difficulty which limits the accuracy of these measurements is the proper handling of the lead wires. For this reason it is usually preferable to measure the tube in some kind of a jig or socket. The socket can be rigidly connected to the bridge and switching key as shown, and measured by itself. A tube is then inserted and the increase in capacitance is taken as the capacitance of the tube. While this introduces a slight error due to the added capacitance from the tube elements to the metal parts of the socket, this is believed to be at least as small as for any other possible arrangement of leads and since some capacitance of this nature is always present when the tube is in use, it can logically be considered as part of the tube capacitance. If an ordinary socket is used it is important that both filament terminals be strapped together when the socket is measured alone or else the socket capacitance will change when a tube is inserted.

If it is desired to make the measurement without the use of a socket, the leads should be rigidly placed in the position which they will occupy when connected to the tube and their capacitance measured. This, however, does not eliminate the error referred to above since the tube elements will still have capacitance to the leads when brought close enough to make the connections.

Results of Measurements

Tables I and II give measurements on several tubes picked at random illustrating the use of the above method. In Table I the capacitance to ground is included in the plate-to-filament

TABLE I

DIRECT CAPACITANCE AND CONDUCTANCE BETWEEN ELEMENTS OF VACUUM TUBES

	Plate to Grid		Plate to Fil. and Ground		Grid to Fil. and Ground	
	C*	G*	C	G	C	G
Measurements on Sockets and Leads						
Socket for 101-D and 101-F Tubes	0.2	0.0000	7.7	0.0042	7.8	0.0042
Socket for 231-D Tubes	0.6	0.0001	2.1	0.0005	2.1	0.0006
Socket for 215-A Tubes	0.8	0.0002	2.7	0.0006	2.9	0.0005
Measurements on Above with Tubes Inserted						
Tube 101-D No. 1	4.9	0.0003	11.3	0.0046	13.0	0.0047
Tube 101-D No. 2	5.2	0.0004	11.3	0.0045	13.0	0.0047
Tube 101-F No. 1	6.2	0.0002	11.5	0.0044	12.3	0.0045
Tube 101-F No. 2	6.3	0.0003	11.6	0.0047	12.2	0.0050
Tube 231-D No. 1	4.3	0.0008	4.4	0.0012	4.5	0.0013
Tube 231-D No. 2	4.2	0.0009	5.1	0.0011	4.8	0.0012
Tube 215-A No. 1	3.7	0.0005	4.6	0.0012	4.7	0.0009
Tube 215-A No. 2	3.5	0.0003	4.1	0.0006	4.4	0.0006
Increase Due to Tubes						
Tube 101-D No. 1	4.7	0.0003	3.6	0.0004	5.2	0.0005
Tube 101-D No. 2	5.0	0.0004	3.6	0.0003	5.2	0.0005
Tube 101-F No. 1	6.0	0.0002	3.8	0.0002	4.5	0.0003
Tube 101-F No. 2	6.1	0.0003	3.9	0.0005	4.4	0.0008
Tube 231-D No. 1	3.7	0.0007	2.3	0.0007	2.4	0.0007
Tube 231-D No. 2	3.6	0.0008	3.0	0.0006	2.7	0.0006
Tube 215-A No. 1	2.9	0.0003	1.9	0.0006	1.8	0.0004
Tube 215-A No. 2	2.7	0.0001	1.4	0.0000	1.5	0.0001

* Capacitance in micromicrofarads and conductance in micromhos.

TABLE II

DIRECT CAPACITANCE AND CONDUCTANCE BETWEEN ELEMENTS OF VACUUM TUBES

	Plate to Filament		Grid to Filament	
	C	G	C	G
Measurements on Sockets and Leads				
Socket for 101-D and 101-F Tubes	1.5	0.0009	1.5	0.0008
Measurements on Above with Tubes Inserted				
Tube 101-D No. 1	3.2	0.0010	5.3	0.0012
Tube 101-D No. 2	3.2	0.0010	5.2	0.0012
Tube 101-F No. 1	3.2	0.0009	5.2	0.0010
Tube 101-F No. 2	3.2	0.0010	5.1	0.0015
Increase Due to Tubes				
Tube 101-D No. 1	1.6	0.0001	3.8	0.0004
Tube 101-D No. 2	1.6	0.0001	3.7	0.0004
Tube 101-F No. 1	1.6	0.0000	3.7	0.0002
Tube 101-F No. 2	1.6	0.0001	3.6	0.0007

and grid-to-filament measurements as would be the case when the tube is used with the filament circuit grounded. Table II gives the corresponding measurements with the ground capacitance eliminated. Comparison shows that 1-½ to 2 micromicrofarads of these capacitances as given in Table I are capacitance to ground. This is due partly to the fact that the socket used has a metal shell which was grounded and which increases the capacitance to ground in the base of the tube. The grid-to-plate capacitance of course is always measured with the ground capacitance eliminated.

The above measurements were made at a frequency of 1000 cycles. It was originally in connection with the development of telephone repeaters and audio-frequency amplifiers that measurements of this type were required. However, since a con-

siderable part of the capacitance is through the vacuum and does not change with frequency, the change of capacitance with frequency which is due to the solid dielectric cannot be a very large proportion of the total. Therefore, audio-frequency capacitance measurements can usually be applied to radio-frequency computations without excessive error. However, when it is desired to take the conductance into account in radio-frequency computations it is desirable to have radio-frequency measurements.

With proper bridge design the same method can of course be applied to radio-frequency measurements, the principal difference being in the transformers and detector. A bridge of this general type but having only one transformer and therefore suitable only for measurements with the *D* corner grounded has been used experimentally at frequencies up to 1500 kc. Measurements of plate-to-filament and ground, and grid-to-filament and ground capacitance were made on this bridge at a frequency of 1000 kc. on some of the tubes of Table I and were found to check the audio-frequency measurements within one- or two-tenths of a micromicrofarad. The corresponding conductance measurements were too uncertain to give any significance to the results. However, the tests indicate that the method is applicable at radio frequencies although, on account of the very small quantities to be measured, great refinement is necessary in the physical construction of the bridge and accessory apparatus.

Discussions on
THE DISTORTIONLESS RECEPTION OF A MODULATED WAVE AND ITS RELATION TO SELECTIVITY
(F. K. Vreeland)*

Henry Shore:† There is one very important factor that Dr. Vreeland has overlooked in his excellent paper. This factor is that the transmitter has the same inherent fault as the ordinary receiver. That is, the higher audio-frequencies are discriminated against much the same as in receiver as shown in Fig. 3 of Dr. Vreeland's paper.

Thus, the same means used to square up the frequency characteristic of the receiver might be used to make the transmitter's characteristic linear. If this were done, considerable power would be sacrificed in the additional circuit. Consequently, it would appear better to exaggerate the peaks of the frequency characteristics shown in Figs. 10, 16, and 17 in order to compensate fully for the non-linear characteristic of the transmitter.

J. R. Nelson:‡ Dr. Vreeland's explanation of the behavior of his "band selector" is very ingenious and interesting. He gives as one frequency where the reactance will be zero the value determined by the constants of one side of his balanced circuit.

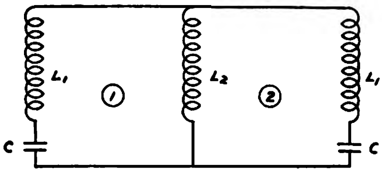

Fig. 1

There are, however, several points regarding the operation that are not clearly brought out in the paper. These are:

The other frequency at which the circuit will have zero reactance; or

The width of the resonance curve, and

The variation of width with frequency.

It was suggested during the discussion after the meeting that this circuit was a tuned coupled circuit. Leaving this question

* Presented at the Annual Convention of the Institute of Radio Engineers, January 9, 1928. Proc. I.R.E., 16, 3; March, 1928.

† Research Engineer, Radio Corporation of America, New York City.

‡ Radio Engineer, E. T. Cunningham Co., Inc., New York City.

aside, however, the circuit may be analyzed mathematically as shown below.

Fig. 1 shows the circuit. This is an electrical network with two meshes, so the network will have two degrees of freedom. Imagine one of the condensers charged and then allowed to discharge through the circuit. Oscillation would occur as the resistances are neglected. Using Kirshoff's first law the equations for meshes (1) and (2) are:

$$I_1\left[P(L_1+L_2)+\frac{1}{PC}\right]+PL_2I_2=0 \tag{1}$$

$$I_2\left[P(L_1+L_2)+\frac{1}{PC}\right]+PL_2I_1=0 \tag{2}$$

where $P=\frac{d}{dt}$ and $\frac{1}{P}=\int dt$

Both sides have the same constants so L will be written for L_1+L_2

$$I_1\left(PL+\frac{1}{PC}\right)+PL_2I_2=0 \tag{3}$$

$$I_2\left(PL+\frac{1}{PC}\right)+PL_2I_1=0 \tag{4}$$

from (4) $I_2=-\dfrac{PL_2I_1}{PL+\dfrac{1}{PC}}$ (5)

Substituting (5) in (3) and clearing of fractions

$$P^4(L^2C^2-L_2^2C)+2P^2LC+1=0 \tag{6}$$

Solving (6) for P^2

$$P^2=-\frac{LC\pm L_2C}{L^2C^2-L_2^2C^2} \tag{7}$$

We are only interested in the steady conditions,

therefore $P=jw$ (8)

$$w^2=\frac{LC\mp L_2C}{L^2C^2-L_2^2C^2} \tag{9}$$

The meshes have a common impedance so that

$$K=\frac{L_2}{\sqrt{L^2}}=\frac{L_2}{L} \tag{10}$$

therefore $L_2=KL$ (11)

$$\omega^2=\frac{LC(1\mp K)}{L^2C^2(1-K^2)}=\frac{1}{LC(1+K)} \text{ or } \frac{1}{LC(1-K)} \tag{12}$$

Let

$$\omega_1=\sqrt{\frac{1}{LC}} \tag{13}$$

Then

$$\omega'=\frac{\omega_1}{\sqrt{1+K}} \tag{14}$$

$$\omega''=\frac{\omega_1}{\sqrt{1-K}} \tag{15}$$

Substituting for k

$$\omega'=\frac{1}{\sqrt{(L_1+L_2)C}\sqrt{\dfrac{2L_2+L_1}{L_1+L_2}}}=\frac{1}{\sqrt{(2L_2+L_1)C}} \tag{16}$$

$$\omega''=\frac{1}{\sqrt{(L_1+L_2)C}\sqrt{\dfrac{L_1}{L_1+L_2}}}=\frac{1}{\sqrt{L_1C}} \tag{17}$$

From this anaylsis we find that one of the frequencies at which this circuit has zero reactance is determined by the constants of one side of the "band selector." The other is determined by $\sqrt{(2L_2+L_1)C}$. Thus it is easily seen that this circuit can be adjusted to any desired band width by changing L_2. Referring to equations (14) and (15) it is seen that as $\sqrt{\mathrm{I}+K}$ and $\sqrt{\mathrm{I}-K}$ are constant, and L_2 is constant, the width of the resonance curve varies with frequency. It would be approximately three times as wide at 1500 kc. as at the 500 kc.

There are several possible methods of using a common reactance which would tend to keep the width of the resonance curve constant independent of frequency. One way would be to use a variable inductance for L_2, which would necessitate, however, another control making three controls for one circuit.

F. K. Vreeland: The point raised by Mr. Shore as to the trimming of side bands in the transmitter is an interesting one.

Mr. Shore in his written discussion has correctly quoted the author's statement at the meeting that the principle of the band selector may be applied to a transmitting system, giving it a band transmission characteristic. It is not clear, however, that this involves a material sacrifice of power. The result can be obtained with substantially the sáme efficiency that is secured in a transmitter using an ordinary tank circuit associated with the antenna.

In Mr. Nelson's discussion, his derivation of the limiting frequencies is perhaps more elegant than that given in the paper, but the result is the same. Thus the last equation on page 12 gives the condition for the limiting frequency F_1. Substituting the values of the circuit constants this becomes

$$x_1 = x_2 = 0 = L_1\omega_1 - \frac{1}{\omega_1 C_1}$$

whence we get the familiar expression

$$\omega_1 = \frac{1}{\sqrt{L_1 C_1}}$$

Similarly, the preceding equation states the condition for the limiting frequency F_2. Substituting the specific values of the reactances x_1, x_2, and x_3 this becomes

$$L_1\omega_2 - \frac{1}{\omega_2 C_1} = -2L_3\omega_2$$

whence we obtain

$$\omega_2 = \frac{1}{\sqrt{(L_1 + 2L_3)C_1}}$$

These results are the same as those given by Mr. Nelson though he has used the symbol L_2 for the inductance L_3.

BROADCAST CONTROL OPERATION*

By

CARL DREHER

(Staff Engineer, National Broadcasting Company, New York City)

Summary—*This paper is limited to a consideration of the audio-frequency elements of a broadcast control system. A two-studio electrically interlocked plant suitable for network operation is described. The methods of specifying and measuring telephonic energy levels, arranging low impedance (as 500-ohm) and bridging apparatus, equalizing lines, and maintaining the audio energy within permissible limits by means of amplifying and attenuating units are described in connection with the specifications of the plant. The co-ordinative and regulative functions of the technical staff of a broadcasting system, the relations of engineering and studio personnel and typical precautions against breaks in program continuity are then discussed.*

FUNDAMENTALLY broadcast operation is radio telephone transmitter operation with a complicated input system, which takes in sound energy from one group of persons, the performers, is operated by a separate group of technical men, and, in addition, is usually tied up with wire telephone lines. This diversity of function introduces many problems of organization and coordination, both electrical and human. Furthermore, the necessity for uninterrupted program service makes it essential to time everything to the second, to leave nothing to chance, and to maintain many delicate adjustments of apparatus and procedure. Thus there is considerable difference between the operation of a radio-telephone transmitter by more or less skilled personnel, for communication purposes only, and the functioning of a broadcasting system.

The technical problems of a broadcasting system may be divided into audio-frequency operation, which includes pick-up of sound energy, the control of telephonic power levels, switching, and line operation; and radio-transmitter operation. This paper will be confined to the audio-frequency elements, including the operation of broadcast studios, the associated control rooms, field pick-up, and some features of network operation. The writer does not wish to claim originality for the material presented. The aim has been to get together a compendium of operational practice as it has developed in various broadcast

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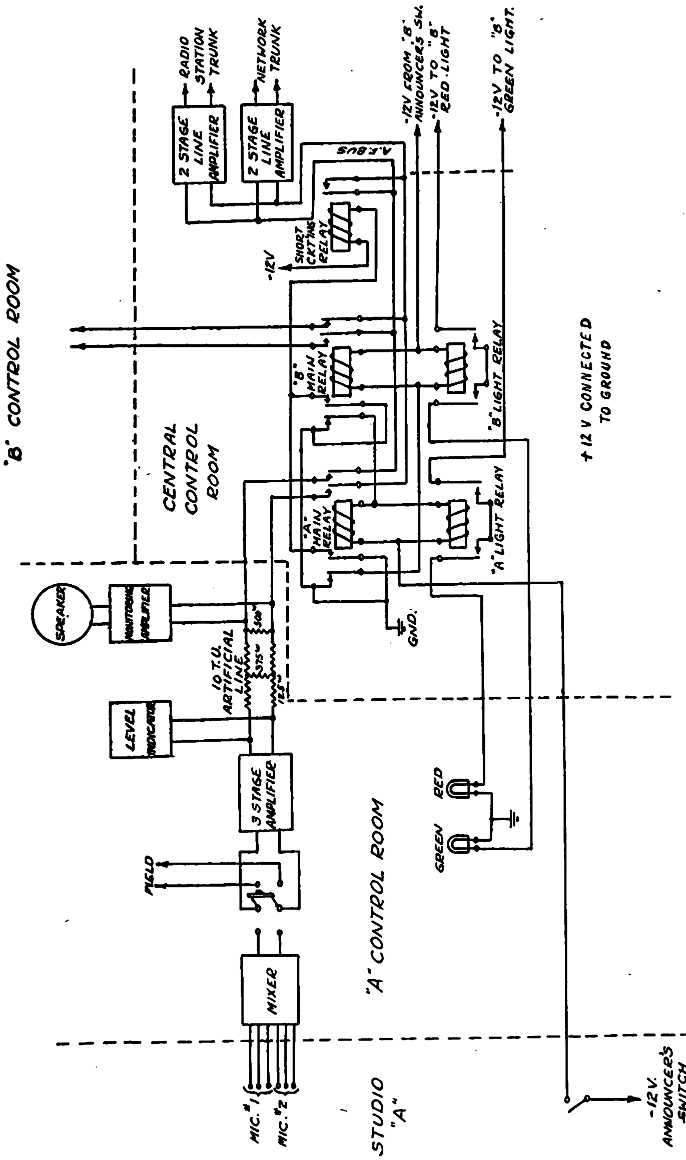
'B' CONTROL ROOM
CENTRAL CONTROL ROOM
'A' CONTROL ROOM
STUDIO "A"
RADIO STATION TRUNK
NETWORK TRUNK
2 STAGE LINE AMPLIFIER
2 STAGE LINE AMPLIFIER
A.F. BUS
SHORT CKT'ING RELAY
-12V
'B' MAIN RELAY
'B' LIGHT RELAY
'A' MAIN RELAY
'A' LIGHT RELAY
-12V FROM 'B' ANNOUNCER'S SW.
-12V TO 'B' RED LIGHT
-12V TO 'B' GREEN LIGHT
+12 V CONNECTED TO GROUND
SPEAKER
MONITORING AMPLIFIER
LEVEL INDICATOR
10 T.U. ARTIFICIAL LINE
500Ω
375Ω
125Ω
GND.
3 STAGE AMPLIFIER
FIELD
MIXER
MIC. #1
MIC. #2
RED
GREEN
-12V. ANNOUNCER'S SWITCH

Fig. 1

stations, and the instruction books of manufacturers of broadcast equipment, the experience of the writer, and the work of his colleagues, have been drawn on with equal freedom.

In general, it will be remembered that considerations of quality or naturalness of reproduction, and efficiency in transmission, require (1) the design of the system to respond more or less impartially to the frequencies within the practical audio band; (2) the matching of impedances of connected elements; and (3) the confining of energy levels between such limits that (a) noise will not become noticeable, (b) cross-talk, either into or from broadcast circuits, may be avoided, and (c) overloading of tubes or other circuit elements will not occur. The theoretical and practical reasons for these precautions are well known in the literature and do not differ fundamentally in broadcast transmission from the same considerations in other branches of the communication arts.

Fig. 1 is a schematic outline of a more or less typical two-studio and control lay-out suitable for simple network operation. The design of the component parts, such as the mixer and amplifiers, will be discussed later. Following the circuit from the left, we see that the input of the first amplifier may be connected either to an outside (field) pick-up point, or to the studio microphones, of which two are shown combined in a mixer. The actual switching would not be accomplished with a knife switch, as shown in the diagram for the sake of simplicity, but by means of a telephone key or plug-jack arrangement in which the transition from one circuit to another is made without a long break. The output of the amplifier is connected to an artificial line which drops the level 10 *TU* (telephonic transmission units).[1] This line presents

[1] The telephonic transmission unit is merely a convenient means of expressing energy or current ratios logarithmically. The number of *TU* is given by the formula

$$TU = 10 \log_{10} \frac{P_1}{P_2} \tag{1}$$

where P_1 and P_2 are two powers or quantities of energy. If the impedance remains the same it follows that

$$TU = 20 \log_{10} \frac{E_1}{E_2} = 20 \log_{10} \frac{I_1}{I_2} \tag{2}$$

E_1 and E_2 being voltages, I_1 and I_2 currents.

an impedance of 500 ohms in both directions. The amplifier has the same output impedance, while the input impedance is 200 ohms, which is correct for a standard double-button carbon transmitter, or a number of such microphones combined in a mixer. The matching requirements are not highly critical and a feeding impedance of, say, 130 ohms, which is the value for one type of mixer, is allowable. Continuing with the main circuit, we note that a level indicating device is bridged across the input of the artificial line. This indicator is designed to give correct readings when connected across a 500-ohm circuit of the type shown. The output of the artificial line feeds a monitoring amplifier, but the main circuit passes through a switching system to the input of two line amplifiers containing two stages of amplification each. One of these amplifiers feeds the local radio station, while the other transmits telephone currents to a network. The purpose of splitting the circuit by means of one-way repeaters at this point is to permit local announcements to be made by short-circuiting the network amplifier, thus confining whatever is said in the studio to the radio station which remains connected. The function of the artificial line, which may seem an unnecessary loss device between the preliminary and line amplifiers, is to provide an intermediate low-impedance circuit with constant level, suitable for measurement purposes, and unaffected by the addition of ordinary bridging impedances.

The electrical operation of the interlocked studios may be understood from an inspection of the relay circuits. In Fig. 1 the details are shown only for Studio *A*, the circuits of Studio *B*

While it is necessary to calculate the *TU* loss or gain in each section of a circuit from (1) or (2), which in turn requires full engineering data secured in the usual way and in the usual fundamental units, once the *TU* figures have been obtained they may be added for the over-all attenuation or "gain." This is not only more convenient than multiplication and division when such computations are a matter of daily employment, but as both the attenuation of lines and the subjective sensation of loudness in the ear are logarithmic functions, the *TU* calculation is a direct quantitative reflection of a practical physical condition in these and other cases.

Plus and minus in the *TU* system are reckoned from an arbitrary "zero level" of 10 milliwatts, which is assumed as the output of a standard commercial telephone transmitter on peaks of normal speech. This corresponds to a current of 4.47 milliamperes in a 500-ohm circuit. Thus a power of 100 milliwatts would be expressed as a telephonic level of plus 10 *TU* while 1 milliwatt corresponds to a level of minus 10 *TU*. One *TU* also corresponds approximately to the loss along one mile of standard No. 19 A.W.G. telephone cable.

being precisely similar. The output of the 10 *TU* artificial line goes to two fixed contact points of a telephone relay. The winding of this relay is supplied with 12 volts when the announcer's switch in the studio is closed. If the *B* main relay is open, the other terminal of the *A* relay winding is connected to the positive side of the 12-volt battery, which is grounded. The same action supplies current to a red light in the *A* studio or control room, visible to both operator and announcer, warning both of them that their studio is connected through to the line amplifiers. If the *B* studio is in use, then it is impossible to connect the *A* studio to the audio-frequency bus which leads to the line amplifiers, since the *B* relay keeps the *A* relay coil from being energized. Thus there is no possibility of putting two studios on the air simultaneously. The studio which is not in use receives a green light from the studio which is in operation. If neither studio is in use both lights are dark. In this condition, also, the audio-frequency bus supplying the line amplifiers is short-circuited by an auxiliary relay, as shown in Fig. 1. Were this not done the input of the line amplifiers would be open when neither studio was "on the air," with the result that if these amplifiers had their filaments lit cross-talk pick-up would be likely to go out on the lines.

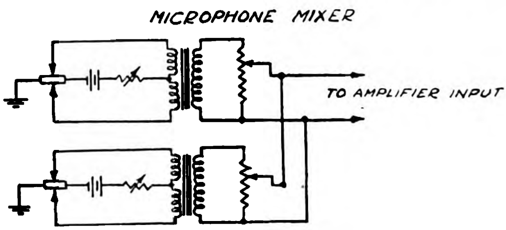

Fig. 2

Fig. 2 shows the circuits of a microphone mixer which allows the outputs of two microphones to be combined in any desired proportion. In the case of standard carbon transmitters, the a-c. impedance being 200 ohms, one-to-one repeating coils with windings of the same impedance are used to match the transmitters, and to provide a mixing circuit free from any d-c. flow which would give rise to noise when variations were made. The potentiometers are of 400 ohms total resistance. The combination of several such elements works into the 200-ohm input of the first amplifier.

As for level considerations, a typical instance would start with the output of the mixer some 50 *TU* below zero level, which may be raised by the first amplifier to slightly above zero level, say plus 2 *TU*. (Peaks of modulation are always understood in such expressions.) In other words, the "gain" of the first amplifier is commonly 50 *TU* in average broadcasting. This corresponds to an energy amplification of 100,000 times, or a voltage amplification of 316. This first amplifier, which may be termed the microphone amplifier, generally uses three tubes, impedance- or resistance-coupled, to produce this amplification. The available gain may actually be 80 *TU*, which, by means of a potentiometer gain control on the input of the second stage, combined with a tap arrangement on the secondary of the input transformer, is variable in steps of about 3 *TU* from 20 *TU* to the maximum. The first tube may have an amplification constant of 30, while the two following triodes have an amplification constant of 6 or 7. The safe undistorted output is plus 10–12 *TU*, corresponding in the case of the lower figure to 100 milliwatts of audio energy. Such a tube will have an oscillator rating in the neighborhood of 5 watts.

The plus 2 *TU* level in the output of the first amplifier is dropped to about minus 8 *TU* by the attenuation network. Such a network has the property of presenting any desired impedance, within practical limits, in either direction, while introducing a loss of the desired magnitude. If an attempt were made to drop a certain number of transmission units by means of a simple shunt resistance, the impedance requirements looking forward and backward could not be met. The design of such a network as the H-form shown is, within limits, simply a problem of calculating currents by Kirchhoff's Laws and expressing the results in telephonic units. However, the network design cannot be relied upon unless definite terminal conditions are met; for example, the transformers must be flat within given frequency limits, and reasonably close to ideal design in such matters as magnetic leakage and open-circuit impedance. The relations are of course reciprocal, so that unless the artificial network design is correct the terminating transformers will also lose quality. It will be noted, in the case of the 10 *TU* line shown in Fig. 1, that a 500-ohm resistance must be connected across the output in order to carry out the design.

The line output level of minus 8 *TU* is suitable for feeding the monitoring amplifier, which is a bridging amplifier, that is, one with a high-impedance input, in this case about 12,000 ohms, suitable for connection across a 500-ohm circuit without affecting the quality or telephonic energy level of the latter. In this case, by means of two stages of amplification the level is raised about 20 *TU* to plus 12 *TU*, which is sufficient to drive a cone speaker with a good monitoring signal.

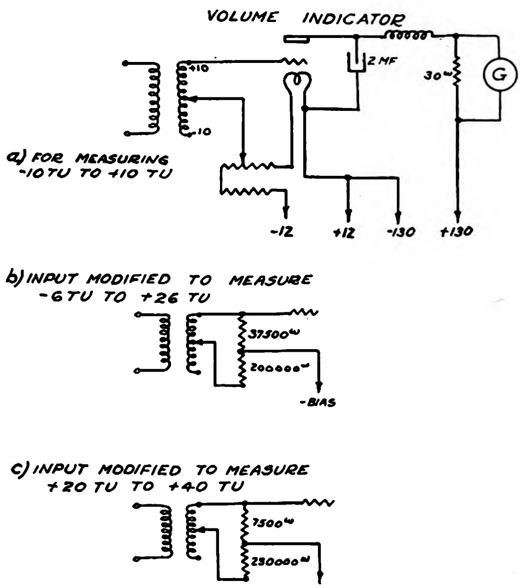

Fig. 3

In the main broadcast circuit the minus 8 *TU* level is supplied to the audio-frequency bus and carried to the two-line amplifiers, which are similar to the monitoring amplifier. These are also used with the 12,000-ohm input, so that a considerable number may be bridged across the bus without affecting each other. The minus 8 *TU* is stepped up in each of these amplifiers to plus 2 *TU*, which is the standard for input to the lines when conditions are normal. The output of the line amplifiers is 500 ohms, to match the cable impedance. In addition to the reason which has already been given for the use of separate line amplifiers, in that they afford independent control of the outgoing circuits, it is also found a convenient arrangement when a large number of studios, each with an associated control room, are handled through a central

control office where most of the functions of measurement and outside communication are concentrated.

A description of a level indicator known as the Western Electric 518-B type may be of interest. The circuits of this instrument, which is substantially a thermionic voltmeter used to measure audio-frequency alternating voltages, are shown in Fig. 3. The instrument is of the bridging type, with an input impedance of about 10,000 ohms. The grid of the high-μ tube is biased negatively so far that the plate current is reduced to a reading of 5 out of 60 scale divisions on a certain type of d-c. galvanometer, the terminals of which are shunted by 30 ohms. This corresponds to a plate current of about 0.2 milliampere. Under these conditions the tube rectifies. When the rectified current, smoothed out to syllable frequency by the combination of inductance and capacity in the plate circuit, causes peaks of 30 scale divisions, corresponding to about 1.12 milliamperes plate current, the tap on the secondary of the input transformer and the value of the potentiometer shunt across it may be read to express the telephonic level of the 500-ohm circuit across which the instrument is bridged. A key controls steps of 0, 16, and 30 *TU*, while an auxiliary tap switch may be set to a value between minus 10 and plus 10 *TU*. The latter is the coil tap arrangement shown in Fig. 3, while the large steps are determined by the setting of the grid voltage divider. The instrument measures from minus 10 to plus 40 *TU*, is itself flat to within 1 *TU* between 100 and 5,000 cycles, introduces a slight and correctable loss at various frequencies, and may be used to measure voltages or currents as well as *TU* levels. While its indications are variable with the wave form of the voice or musical currents under measurement, so that some percussion instruments, for example, do not cause the galvanometer needle to swing as high as rounder peaks of the same amplitude, thus introducing the possibility of misleading indications, it is a valuable visual check in broadcast stations. Control methods based on hearing alone would be extremely primitive; the ear has no bright possibilities as a precision instrument. Of course a visual indicator based on the use of a photo-electric or other relatively inertialess response instrument would dispose of the ballistic and form factor complexities of a tube-filter-d-c. galvanometer arrangement, but no such equipment has yet been devised in a shape suitable for application outside of the laboratory.

Fig. 4 is a sketch of the arrangement employed for field broadcasting. The field amplifier is often easily portable, with dry-cell tubes of limited output capacity, say plus 2 *TU*. The amplifier is then operated at a peak level of minus 4 *TU*, affording 6 *TU* overload margin. This is convenient for volume indicator measurement (the field-amplifier also contains a volume indicator) but higher than necessary for a quiet line, so that a 10 *TU* artificial line is inserted, reducing the level at the line terminals to about minus 14 *TU*. If the line has a 1000-cycle equivalent of 6 *TU*, meaning that it introduces a loss of 6 *TU* at the mean speech frequency, the telephone currents will reach the broadcast station at minus 20 *TU*. As the volume at the receivers must not change

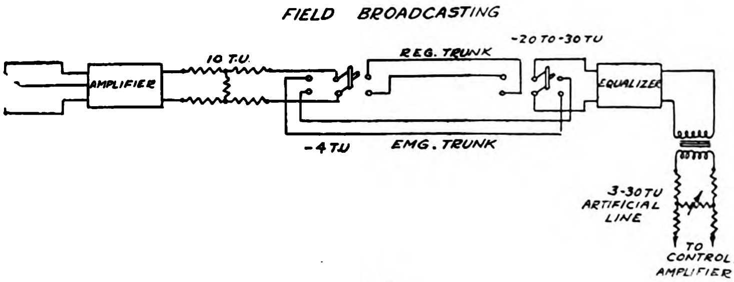

Fig. 4

appreciably during field-studio change-over operations, this is still too high for input to the station amplifier, which, it will be recalled, takes energy from the studio at about minus 50 *TU*. An artificial line with attenuation variable between 3 and 30 *TU* makes up the difference. By this means the change from studio to field pick-up, or vice versa, may be made without an abrupt change in level.

So far we have not discussed in any detail frequency characteristics of broadcast circuits. The presence of the equalizer in Fig. 4 brings up this consideration. In general, broadcast circuits are now designed to be "flat" between 100 and 5000 cycles per second, within 1 *TU*. That is, the over-all transmissi n characteristic from microphone to antenna should be horizontal for all audio frequencies within this band. A line, because of its distributed capacity (0.054 μfd between wires per mile for standard No. 19 A.W.G. cable) will manifest a progressively higher attenuation for the higher frequencies. Fig. 5-A shows this characteristic. By means of an equalizer, which is a network possessing a

compensating characteristic (Fig. 5-B), flat where the line is flat and with losses decreasing where the line losses increase with frequency, a resultant may be secured which brings up the relative strength of the high notes and smooths out the over-all characteristic for the essential band of frequencies (Fig. 5-C).

The equalizer, shown schematically in Fig. 6, has a variable impedance for different frequencies. It consists of a parallel circuit resonant to approximately the upper limit of frequency to which it is desired to extend the line characteristic, in series with a variable resistance. The function of the resistance is to control the extent to which the parallel circuit will affect the

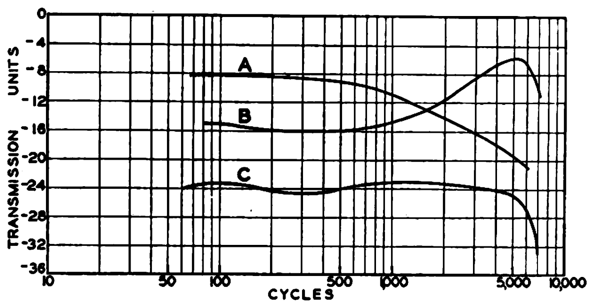

Fig. 5—*A*, Attenuation Characteristic of Cable with 11 *TU* Equivalent at 1000 cycles. *B*, Characteristic of Shunt Equalizer, $R = 10\ \Omega$. *C*, Characteristic of Cable and Equalizer.

transmission characteristic of the whole circuit. For low frequencies the inductive reactance of the coil is negligible (3.14 ohms at 100 cycles for a 5-millihenry coil) so that the value of the shunt is practically determined by the resistance in series. As the frequency rises the inductive reactance of the network increases proportionately, thus increasing the shunt impedance for the higher notes. In the meantime the capacitive reactance of the condenser has been decreasing inversely as the frequency, and at resonance the shunt impedance is a maximum. It drops off again at higher frequencey, causing a cut-off. The loss introduced by such a network is about 10 *TU* at the mean speech frequency (1,000 cycles). With this type of equalizer it is not feasible to correct the characteristic of a line having an equivalent by itself above 12 *TU* at 1,000 cycles.

Now that an outline of the technical basis of broadcast operation has been given, we may consider in some detail the actual procedure whereby programs are put on the air.

The function of the control operators, whether in the field or at the studio, is partly coordinative, as in connection with inter-studio contact and switching, and partly regulative, in that it is found necessary to compress the natural volume variation of speech and music, which may be as high in some cases as 60 *TU*, into a compass of about 40 *TU*, if overloading is to be avoided on the one hand and noise interference on the other. The operator makes up this 20 *TU* difference, in exteme cases, by bringing up his gain control carefully on low passages. Some vocal artists who have adapted their renditions to the requirements of broadcast transmission take care of this themselves by avoiding extreme pianissimos or by swaying back and forth as they sing, approach-

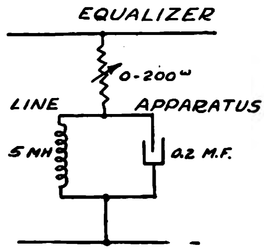

Fig. 6

ing the microphone during pianissimo portions and withdrawing during fortissimos. The former procedure may also be followed by orchestras. The rule is for the operator to handle the gain control as little as possible, but to regulate it when necessary to avoid overloading or the loss of low passages. The volume indicator galvanometer should flicker slightly during the peaks of low passages and rise to the maximum of 30–40 scale divisions during the loudest intervals. Gain regulation must be confined to one place, which is logically at the point of control nearest the origin of the program. The field operator therefore assumes the function of changing the level when necessary, in the case of field programs, the studio operator in the case of studio programs, and the transmitter operator only in the event of serious line irregularities or careless operation which may endanger the radiation of the program at his end.

With regard to the placing of performers in the studio for the best musical balance there is some difference of opinion as to the proper arrangements. In general the non-technical studio staff wishes to place the musicians conveniently (for them) and to

move microphones freely. The engineers, on the contrary, prefer a fixed position for the microphones, neccessitating the grouping of the musicians to secure the best musical balance. On this basis the microphone position is fixed according to the acoustic characteristics of the room. In studios which have not been highly damped it is frequently found that standing waves set up at certain frequencies between reflecting surfaces manifest themselves in their various interferences as rattling sounds following an initial impulse. It is possible, by placing a small rug on the floor near a wall drape, to form a space relatively free from such acoustic disturbances, in which the microphone will pick up a program with greater freedom from disturbing transients and distortions in reproduction. This amounts to stating that optimum microphone placing is a function of the studio characteristics and should be left to the judgment of the electro-acoustic experts, not to that of musicians, who, as a class are lacking in scientific qualifications. Musical balance then becomes a problem in placing instruments with reference to the fixed microphones and standardizing on the best positions. This responsibility may devolve on a musician especially delegated to the task, or on the announcer if he has musical training, or on a committee of musicians and musically-experienced technical men capable of listening critically and objectively to loud-speaker reproduction. It is often helpful to allow the conductor of an orchestra to listen, during rehearsals, to the monitoring reproduction, either while his men play without him or under the baton of an assistant. The questions of orchestral balance involve many factors of musical taste, imagination, individual auditory characteristics, and imponderables, which make agreement difficult at best. The problems involved are complex and their full discussion would require a separate paper on the acoustic and musical principles underlying them.

In a broadcast station all program matters are laid out beforehand and printed schedules detailing the artists, announcers and announcements, selections, timing, and studio arrangements are distributed to all personnel concerned. The chances of a slip-up are further reduced by the fact that all program "features" are carefully rehearsed and timed beforehand. While this system does not contribute to spontaneity it has been found the only means of running off a complicated program with dispatch and reliability, especially in chain broadcasting. The function of the operating

personnel, under such arrangements, is reduced, save in emergencies, to following routine previously established.

Communication between studios is maintained by means of telephone systems. The operator in the control room associated with each studio, and seated within sight through a double glass window, is in touch with the other studio by means of a breast transmitter and single head-band receiver. He is thus in a position to converse with the other operator while continuing to monitor the program going out through his own studio. It is his duty to keep the other operator informed of the progress of the program and to warn him some minutes before a change from one studio to the other is due. Generally head-receiver facilities for listening to the program in another studio are also provided for the announcers, and in the more intricate set-ups of chain broadcasting it is necessary to devise complicated systems of mechanical switching whereby the announcer picks up his station on visual signal from the control room, by pressing buttons which actuate telephone relays and make the necessary circuit changes, which are, however, supervised by the operators, who sit before similar control boxes and are in a position to correct switching errors made by the announcers.

The preliminary procedure of field broadcasting gives a good idea of the precautions taken to prevent breaks in program continuity. Generally two broadcast pairs are provided, in addition to an order pair for speech communication only. The routine is as follows:

(1) The field operator, having set up his microphones and amplifier, calls in on the order pair one hour before program time and talks to the control operator at the station.

(2) The field operator tests all microphones by talking into them with the central operator listening.

(3) The field operator sends test talk or preliminary program material over both regular and emergency broadcast circuits.

(4) The station operator raises the gain of his amplifier 20 *TU* and listens closely for cross-talk from the order wire in his monitoring speaker, if this is available, while the field operator talks on the order wire, the input to the field amplifier having been cut off.

(5) The field operator synchronizes his watch with the station operator, who takes time from a master clock system.

(6) Ten minutes before program time the field operator sends room noise or preliminary program to the station for check of continuity of the broadcast circuit. This is kept on to within two minutes of program time.

(7) At program time the field operator is told over the order wire, "Take it away," ("It" referring to the program) immediately after the broadcast trunk has been connected to the station amplifier input. He gives the signal to his announcer, who is generally within reach of a hand signal, and the remote program starts. Communication is then maintained throughout the program by the two operators for the purpose of criticism of quality and the effecting of any necessary changes.

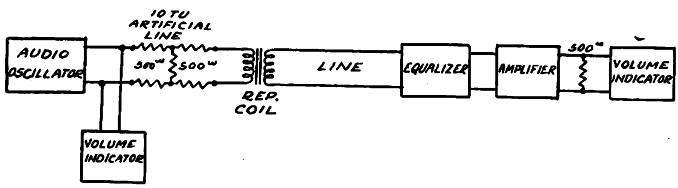

Fig. 7

All such circuits have been previously equalized by means of an audio oscillator sending out tones at a known frequency and level, with an amplifier and volume indicator showing the levels received at the station end. Thus the proper equalizer settings for each line are known and the compensating network is set at this value before or during the above tests. The circuit for the equalization of a wire line shown in Fig. 7, will hardly require comment after the previous description of the functions of the various parts. In general, after the initial run, it is only necessary to check the characteristic daily at three points, say 100, 1,000, and 5,000 cycles. These tests are, of course, in addition to the usual wire chief's d-c. tests for lack of continuity, crosses, grounds, or other defects.

In chain broadcasting similar procedures are followed, the principal difference being that contact between the originating station and the chain is maintained by telegraph. The originating station controls procedure entirely, since obviously with a multiplicity of stations receiving a program unity can be secured only by such a system. The method of making local announcements has been described previously. The fifteen-second intervals in the program left for the announcements are indicated to the chain

stations by telegraph a sufficient time before each pre-arranged gap. The individual stations then cut the line input to their amplifiers and turn over to their local microphones, scurrying back to the chain before the fifteen seconds are up. Test tones are sent out by the head station to the network and the volume indicator readings at the points of reception, telegraphed back to the key station, give a necessary check on wire conditions, possible need for re-routing circuits, etc. All the problems of high-quality telephony, as well as specialized broadcasting procedures, are involved. After each program the syndicate stations wire in reports as to technical quality, entertainment value, and the like. A full consideration of such matters does not fall within the scope of a paper of this type.

I wish to acknowledge the valuable assistance of Mr. R. M. Morris of the National Broadcasting Company in supplying much of the technical detail included in this paper.

THE LORENTZ RECIPROCITY THEOREM FOR ELECTRIC WAVES†

AMONG the tools of thought and artifices by which man forces his mind to give him better service, perhaps the most intensely useful are the simple mathematical rules of inversion known as *reciprocity theorems*. A number of these have been invented and have become of classical importance in almost every field of physics in which the phenomena can be described by means of linear equations. In the field of optics we have the celebrated theorem of Helmholtz (restated by Schuster) concerning the reversibility of light rays, in dynamics the theorem of Rayleigh (recently restated by G. W. Pierce and generalized by J. R. Carson for electric networks), in sound that of Helmholtz, in pure mathematics the theorems concerning the reversal of the modulus and argument of elliptic-integrals, in electro-statics the extremely clear and useful reciprocity theorem due to Green; and now concerning the radiation and reception of electric waves by wireless antennas, we have a reciprocity theorem due to H. A. Lorentz. The following résumé of the theorem is based upon a recent article by A. Sommerfeld and will be of interest to radio engineers. The theorem itself has recently been applied to the calculation of the distribution of radiation about a transmitting antenna erected over an imperfect earth by T. L. Eckersley, and to the equivalent problem of the reception of waves arriving from various altitudinal angles, by L. Bouthillon. As to its history Sommerfeld remarks that the theorem was proved in a dissertation (Munich, 1925) by H. Pfang, but it was subsequently pointed out by M. von Laue that it was really of much greater antiquity, having been clearly stated and proved by H. A. Lorentz[1] thirty years before.

* Original Manuscript Received by the Institute, February 3, 1928.

† A. Sommerfeld: *Zeit. für Hochfrequenztechnik*, **26**, 93, 1925; T. L. Eckersley: *Proc. Wireless Sect.*, I. E. E., **2**, 85, June, 1927; L. Bouthillon: *L'Onde Electrique*, **6**, 533, 1927.

[1] H. A. Lorentz: *Amst. Akad. van Wetenshappen*, **4**, 176, 1895.

1. Statement of the Lorentz Reciprocity Theorem: *If A_1 and A_2 are two antennas situated at O_1 and O_2 respectively and having arbitrary orientations, and signals are first sent from A_1 and received by A_2, and then sent with the same average power from A_2 and received by A_1, then the intensity and phase of the electric field at the receiver A_1 will be equal to that previously produced at A_2, regardless of the electrical properties and geometry of the intervening media (water, land, or combinations of these, stratified or otherwise inhomogeneous atmosphere, of any degree of ionization, etc.) and the forms of the antennas.*

Sommerfeld gives a detailed proof of this theorem for both electrical and magnetic antennas which the reader may wish to consult for further details.

2. Conditions under which the Theorem is Valid: Sommerfeld gives a critical summary of the conditions under which Lorentz's original proof is valid, considering particularly questions (e.g., the state of the upper atmosphere) which have become important in recent years. In the first place no difficulty is presented by its extension to anisotropic media of which the constitutive properties (conductivity σ, and dielectric constant, ϵ) may be different in different directions, provided only that the determinant of these quantities is symmetrical, e.g., $\epsilon_{ik}=\epsilon_{ki}$, and so forth. The proof fails completely, however, when the constitutive relations $D=\epsilon E$, $B=\mu H$, $I=\sigma E$, are non-linear, or when dielectric and magnetic hysteresis are present. Considerable interest attaches to the question of propagation through an ionized medium, but here again it is rather apparent that the validity is unchanged so long as the total convection current remains proportional to the electric force; this would be expected to be the case at all important altitudes. (The motion of ions in the direction of the wave-normal due to "radiation pressure" is not considered).

The theorem fails in a very important case, viz. when the waves are propagated in an ionized medium in the presence of the earth's magnetic field. In these circumstances "gyroscopic" terms are introduced in the motion. As in the Faraday effect in optics a rotation of the plane of polarization occurs and the forward and backward progress of the ray is non-reciprocal. Mathematically the σ's and ϵ's are now partly of an anisotropic character, for example $I_k=\sum_i \sigma_{ik}E_i$, where the σ_{ik}'s are now proportional

to the magnetic field. The structure of ϵ may be represented by[2]

$$(\epsilon)=\begin{vmatrix} \epsilon_1 & -j\alpha & 0 \\ j\alpha & \epsilon_1 & 0 \\ 0 & 0 & \epsilon_2 \end{vmatrix}$$

which is no longer symmetric ($\epsilon_{ik}=\epsilon_{ki}$) as required above, but skew ($\epsilon_{ik}=-\epsilon_{ki}$).

3. Example of Application of the Theorem. Energy Distribution about a Transmitting Antenna over Imperfect Earth: An instructive example of the application of this theorem has been given by T. L. Eckersley, who has considered the distribution of energy about a transmitting antenna when the earth conductivity is finite. The idealized case of an antenna over *perfect* earth has been discussed by van der Pol, Jr. and others [3] and presents a relatively elementary problem. When, however, the earth is taken to be a dielectric-conductor the ordinary theory of images is no longer available for the simplification of the mental work. This more complicated case was rigorously investigated by A. Sommerfeld[4] in a much neglected paper; an extension of this analysis to the case of the horizontal dipol was made by his pupil, H. von Hoerschelmann. Broadly speaking Sommerfeld considers the field produced on the horizon by a vertical dipol at a negligible distance above the surface of a flat dielectric-conducting earth. The results of the somewhat complicated mathematical analysis are quite amenable to calculation when expressed in simple series and in terms of the so-called "numerical distance". An excellent summary of them, with specimen calculations, has been given by Smith-Rose and Barfield[5] which the non-mathematical reader will find especially entertaining and useful.

Sommerfeld's formulas may be applied to the problem of finding the polar distribution of energy about an antenna of finite length; it simply suffices to consider the antenna as a series of

[2] See René Mésny; "Les Ondes Electriques Courtes," p. 44, Paris, 1926.

[3] Balth, van der Pol, Jr., *Proc.* Roy. Soc., 29, 269, 1917. H. Chireix, *Radio-Electricite,* 5, 65, July 1924. Stuart Ballantine, Proc. I. R. E., 12, 833, December 1924.

[4] A. Sommerfeld; *Jahr. der draht Tel. und Tel.*, 4, 157, 1910. H. von Hoerschelmann; *Jahr. der draht Tel. und Tel.*, 5, 188, 1911. T. L. Echersley; previous citation, p. 119, App. I.

[5] R. L. Smith-Rose and R. H. Barfield; *Proc.* Wireless Sect., I. E. E., 1, 182, September 1926.

electric dipols of moments graduated in accordance with the assumed current distribution, and to integrate over the string of dipols. In the case of a perfect earth we represent the effect of reflection by integrating also over a *positive* image A' (current in the same phase as that of the antenna) of the antenna A below the earth's surface. Eckersley first shows by means of Sommerfeld's analysis that over the actual earth the field horizon at large distances is that due to the antenna plus a *negative* image, i.e., one in which the current is 180 deg. out of phase with that of A. This result and the actual calculation of the polar diagram for the antenna's radiation may be checked and obtained more simply by means of the reciprocity theorem as follows:

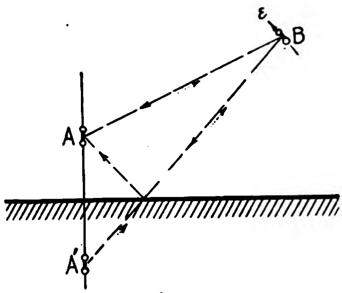

Fig. 1

With reference to Fig. 1, A represents a radiating dipol whose field at a distant point B it is desired to calculate. The wave from A strikes the earth and is reflected, but since A is so close to the earth the wave is certainly spherical when it strikes the earth and it is not at all obvious how the reflection is to be calculated. However, according to the reciprocity theorem the field at B due to A is precisely equal to that at A due to B. The wave from B is, on the other hand, very nearly plane by the time it reaches the reflection point on the earth, and its reflection can be readily calculated by means of the classical Drude equations for a *plane* wave. The total field at A is of course the sum of the direct and reflected waves. The accuracy of the application depends upon the separation of A and B; Eckersley suggests that the results will be accurate provided the distance is large compared with the "numerical distance".

Without going into the details of this calculation the results may be sketched as in Fig. 2. The solid curve represents the variation with the zenith angle ϑ, of the electric force about a

vertical dipol at the earth's surface, assuming infinite conductivity. The effect of decreasing the conductivity is to decrease E at small earth angles as shown by the dotted curve.

Eckersley does not actually carry through the computation of the polar diagrams, but in considering a number of antenna types, suggests that in general the actual curve will follow the curve for perfect earth for polar angles up to about 80 deg. and thereafter approach more closely the curve calculated for a *negative* image. Bouthillon (loc. cit.), on the other hand, carefully calculates diagrams for the cases of sea-water, wet earth and dry earth, rocks, etc., for a receiving antenna upon which are incident rays making various angles with the earth's surface.

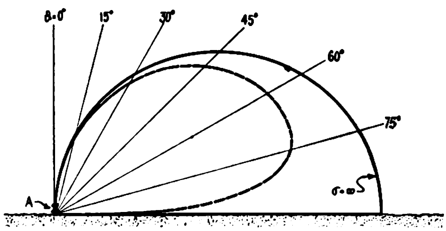

Fig. 2—Variation of Electric Field About Vertical Dipol near Earth's Surface for Cases of Perfectly Conducting Earth ($\varepsilon = 10$; $\sigma = 5 \times 10^7$) at about 70 meters wavelength.

According to the reciprocity theorem these diagrams are of course equally appropriate for the radiation from a transmitting antenna of the same form. Unfortunately Bouthillon's diagrams are not suitable for reproduction.

Figs. 2 and 3 are included to illustrate the actual form of the distribution curves for two representative cases, and were selected from some cases which had been computed for me by Lieut. Raymond Asserson.[6] Fig. 2 represents the case of a vertical dipol at the earth's surface; in Fig. 3 the dipol is elevated above the surface by a distance equal to $\lambda/2$. The assumed earth constants were as follows: $\epsilon - j4\pi\sigma/\omega = 10 - j10$. These values appeared to be the best compromise between the somewhat divergent data of Zenneck, Hack, Loewry, and Smith-Rose and Barfield for ordinary earth at wavelengths between 15 and 150 meters.

The effect of increasing the conductivity is to restore the radiation (and reception) at low angles with the earth, until for

[6] These computations are considerably facilitated by means of Pierce's tables of f and g functions; *Proc.* Amer. Acad., 57, 175, April 1922.

perfect conductivity the curve becomes discontinuous; then the ray along the surface may be mathematically zero, but finite for the slightest physical elevation. The idea is somewhat strange but not hopeless; in the case of sea-water at short wavelengths, for example, a hyper-critic would probably drown himself trying to observe experimentally the low intensities predicted for almost glancing incidence.[7]

It would appear that at distances so great that the ground wave has become very much attenuated a notable increase in signal strength should be observed with increasing elevation aside from any magneto-ionic effects of the upper atmosphere. It would be interesting to check this experimentally by means of a receiver located in an airship, using waves of such short length that the

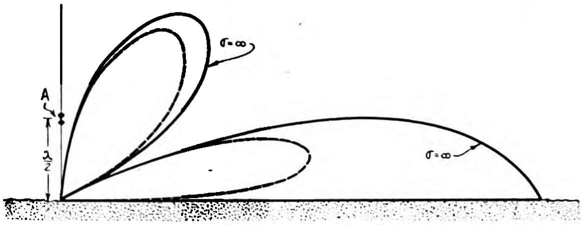

Fig. 3—Variation of Electric Field about Vertical Dipol at a Height of $\lambda/2$ for Cases of Perfect Earth (solid curve) and Actual Earth (dotted curve).

distances could be small and the effect of the bulge of the earth thereby avoided. The results would also suggest that communication at short wavelengths between airplanes might be feasible at medium distances when communication between ground stations over the same distances might be impossible due to the absorption of the ground wave. In other words the extent of the so-called skip-distance zone should be considerably diminished in the case of inter-aircraft signalling due to the enlargement of the ground wave range. This idea appears to me to account for the experimental results recently reported by H. Fassbender[8], who states that in aircraft communication a clearly defined skip-distance is not observed.

Another practical consequence of these calculations, which was pointed out by Eckersley, is the advantage for ground communication of increasing the height of short wave antennas.

[7] The corresponding acoustic case is interesting. Prof. G. W. Pierce recently related to me that on a lake in New Hampshire he was able, by placing his ear close to the water surface, to hear with considerable intensity the voice of a man speaking in a boat three miles away. While destructive reflection is also predicted in this case, the phenomena is not simple and Rayleigh surface waves in the water may also play an important part on the propagation.

[8] H. Fassbender: *Zeit. für Hoch.*, 30, 176, 1927.

BOOK REVIEWS

The Cable and Wireless Communications of the World, BY F. J. BROWN. Published by Isaac Pitman and Sons, London and New York. Price $2.25.

The author of this book was formerly Assistant Secretary to the British Post Office in charge of cables and wireless. He is now Director of the International Cable Companies' Association. The first third of the book contains a brief historical and geographical résumé of the existing cable systems of the world together with a description of the general aspects of manufacture and laying of cables and the technique of cable communication.

Following this is a discussion of the relations between the cables and radio in the communication business of the world from which the conclusion is drawn that up until about the present time cable and radio costs were such that there has been no great inherent difference between the two on that basis. It is pointed out, however, that only the future can determine what will be the change in relationship, if any, brought about by the development of the new high-speed cables, (for example, the permalloy cable), and the increasing use of short waves in radio.

One chapter is devoted to a discussion of the relationships between governments and private agencies in the operation of communication services. Reference is here made to the International Telegraph and Radiotelegraph Conventions.

Several chapters touch on financial questions, such as capital investment, depreciation and rate of return on both cable and radio systems and the rates charged for various classes of telegraphic service.

The book closes with a chapter devoted to international broadcasting. Moreover, the book is written in a style which makes it very easy to read.

The Theory and Practice of Radio Frequency Measurements. BY E. B. MOULLIN. Published by J. B. Lippincott Company, 277 South 6th St., Philadelphia, Penna. First edition, 278 pages and 134 illustrations.

This publication is offered as a handbook for the laboratory and a textbook for advanced students. It consists for the most part of a collection of the best-known methods of measuring the fundamental characteristics of radio-frequency apparatus and systems. Beginning with a fairly complete description of the vacuum-tube oscillator as a source of testing power, the author discusses in

separate chapters the measurement of potential difference and current, frequency, resistance, capacity, inductance, antenna characteristics, and finally, radiated fields. The closing chapter contains notes on miscellaneous items such as measurement of the harmonics of a generator, transformer equations, rectification with a heterodyne, etc.

Alternative methods are generally given, with those preferred by the author receiving the more complete treatment. Sufficient explanatory matter is presented, however, to permit the reader to make an independent choice. The chapters on potential and current and on frequency measurements are particularly good; the only important exceptions to this that have been noted are that in the former there is a highly unsatisfactory discussion of the design of a current transformer and in the latter the important use of the phonic wheel in absolute measurements is omitted.

Resonance methods of measuring capacity and inductance by means of indicating instruments are quite satisfactorily treated. No adequate discussion of null or bridge methods is given, these being briefly dismissed by the author without due consideration of their merits. Similarly, practically nothing is said regarding the use of electrostatic shielding. The latter being almost essential to accurate bridge measurements, one suspects that lack of experience with such shielding may account for the stand taken on the former. Resistance measurements, whether of resistors, inductors, or capacitors are less satisfactorily covered, particularly with reference to suitable standards. Much space is devoted to calculations of skin effect and eddy currents in conductors of a type that would seldom be used in standards of either resistance or inductance and little useful information given as to the resistance or phase angle of convenient laboratory forms.

Taking the book as a whole, it is noted that the various methods are grouped in accordance with the characteristics to be determined rather than according to the principles employed. While this arrangement, of course, makes it very easy to select a method for making a particular test, it does not lend itself so well to the logical development of the subject. Perhaps it is on this account that there is almost no attention given to the interrelations of the methods studied. However, in spite of the adverse comments here made, this contribution to the field of radio engineering fills a definite need.

W. J. SHACKELTON

BIBLIOGRAPHY ON PIEZO-ELECTRICITY*

By

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In response to various requests, there is presented herewith a general bibliography on piezo-electricity and its applications. While the writer hopes that it is fairly complete to the beginning of 1928, still he realizes that some of the literature on the subject has doubtless been overlooked. He would be grateful to any who would call his attention to errors or omissions, since he hopes to publish a supplementary list at some future time.

Part I

Books and Periodical Literature

The following list is arranged alphabetically according to authors. In the cases of a few anonymous publications, however, articles are listed by title or by name of publication. Many references to papers in which piezo-electricity is quite subordinate have been included, where the portion in question seemed of sufficient interest. The English equivalents of foreign titles have been made as literal as possible. For the early literature, which is very voluminous, citation is made only of those publications which are of historical value or otherwise of outstanding importance. More complete references to the early literature on the subject may be found in Nos. 49, 85, 87, 109, 158, 214, and 222 below.

For the convenience of those who make use of the bibliography, the subject-matter has been arbitrarily classified into seven categories, as follows:

A. Fundamentals, theory, and early numerical data.

B. General articles on the piezo-electric resonator and oscillator.

C. Optical and other tests; preparation and mounting of crystals.

D. Luminous effects of vibrating crystals.

E. Standards of frequency and wavemeter calibration.

F. Transmitting apparatus and circuits.

G. Miscellaneous applications.

Appended to each reference are one or more of the above seven key-letters, indicating the general nature of the contents.

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ABBREVIATIONS

Ann. d. Phys. *Annalen der Physik* (Leipzig).
Arch. sc. phys. et nat. *Archives des sciences physiques et naturelles* (Geneva).
Elektr. Nachr.-Techn. *Elektrische Nachrichtentechnik* (Berlin).
Elektrot. ZS. *Elektrotechnische Zeitschrift* (Berlin).
Elektrot. u. Maschinenb. *Elektrotechnik und Maschinenbau* (Vienna).
Exp. W. & W. Eng. *Experimental Wireless & Wireless Engineer* (London).
Göttinger Nachr. *Nachrichten von der Gesellschaft der Wissenschaften zu Göttingen, Math.-Phys. Klasse.*
J. Opt. Soc. Am. *Journal* of the Optical Society of America and *Review of Scientific Instruments.*
J. de phys. *Journal de physique* (Paris).
Phil. Mag. London, Edinburgh and Dublin *Philosophical Magazine* and *Journal of Science.*
Phys. Rev. *Physical Review* (Corning, N. Y.).
Phys. Ber. *Physikalische Berichte* (Braunschweig).
Phys. ZS. *Physikalische Zeitschrift* (Leipzig).
Proc. Amer. Acad. *Proceedings* of the American Academy of Arts and Sciences.
Wiener Ber. *Berichte der Kaiserlichen Akademie der Wissenschaften zu Wien, Math.-Nat. Kl.*
W. World. *Wireless World* (London).
ZS. f. Hochfrequenztechn. *Zeitschrift für Hochfrequenztechnik* (*Jahrbuch der drahtlosen Telegraphie u. Telephonie*) (Berlin).
ZS. f. Instrkde. *Zeitschrift für Instrumentenkunde* (Berlin).
ZS. f. Phys. *Zeitschrift für Physik* (Berlin).
ZS. f. techn. Phys. *Zeitschrift für technische Physik* (Leipzig).

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Part II

Patents

Following the title of each patent is a parenthetical note pointing out the distinguishing features from the point of view of piezo-electricity, without attempting to indicate the full scope of the invention. The date for each U. S. and Canadian patent is that of issue; unless otherwise designated, for British patents the "complete accepted" dates are given, and for French and German patents, the dates of publication. The only exceptions are in a few instances where the desired information was not available.

United States of America

1,347,350. R. W. Moore (General Electric Co.). July 20, 1920. Crystal production. (Growth of crystals, especially Rochelle salt, from solution).

1,353,571. O. Dreibrodt (Elektrochem. Werke). Sept. 21, 1920. Method of and apparatus for forming large crystals. (Growth by continuous circulation of liquid).

1,414,370. A. M. Nicolson (Western Electric Co.). May 2, 1922. Method of making piezo-electrical crystals. (Growth from solution, and mounting).

1,438,965. A. M. Nicolson (Western Electric Co.). December 19, 1922. Piezo-electric device and method of producing the same. (Growth and mounting of Rochelle salt crystals).

1,438,974. E. C. Wente (Western Electric Co.). December 19, 1922. Piezo-electrical voltage indicator. (Deformations of piezo-electric plates cause a mirror to rotate).

1,450,246. W. G. Cady (Radio Corp. of America). April 3, 1923. Piezo-electric resonator. (Frequency standard, filter, coupler).

1,452,933. M. I. Pupin (Westinghouse Co.). April 24, 1923. Selective amplifying apparatus. (Multi-stage tube amplifier with "wave balance"; use in connection with piezo oscillator-receiver for submarine signaling).

1,472,583. W. G. Cady (Radio Corp. of America). October 30, 1923. Method of maintaining electric currents of constant frequency. (Piezo-electric stabilizer and oscillator).

1,495,429. A. M. Nicolson (Western Electric Co.). May 27, 1924. Piezophony. (Use of Rochelle crystals for receiving and transmitting sound waves).

1,525,823. A. M. Nicolson (Western Electric Co.). February 10, 1925. Piezoelectrical transmitter. Same as Canadian Pat. No. 250,631. (Rochelle salt crystal, girdle mounting).

1,526,319. L. W. Chubb (Westinghouse Co.). February 17, 1925. Piezo-electric loud speaker. (Vibrations of crystal control stream of air).

1,559,116. W. A. Marrison (Western Electric Co.). October 27, 1925. Wave generating and modulating system. (Various methods for obtaining two frequencies from one generating circuit, using one or two piezo-electric plates).

1,561,278. M. I. Pupin (Westinghouse Co.). November 10, 1925. Wave signaling system. (For submarine signaling, two piezo-receivers so placed that undesired noises are excluded).

1,562,578. A. M. Nicolson (Western Electric Co.). November 24, 1925. Piezo-electric device and method of producing the same. (Rochelle salt crystal with interior electrode, and lever for attachment to loudspeaker).

1,565,566. R. V. L. Hartley (Western Electric Co.). December 15, 1925. **Translating device.** (To record voice on motion-picture film, polarized light passes through piezo-electric plate).

1,572,773. A. Crossley. February 9, 1926. **Piezo-electric crystal apparatus.** (Mounting and hermetical sealing of crystal).

1,574,302. A. M. Nicolson (Western Electric Co.). February 23, 1926. **Piezo-electric loud speaker.** (Rochelle salt crystal, with mounting for attachment to loudspeaker cone).

1,578,296. A. H. Taylor (Wired Radio). March 30, 1926. **Multifrequency crystal-controlled oscillator.** (Two or more crystal plates stacked one upon another). Same as British Pat. No. 259,174.

1,581,701. A. H. Taylor (Wired Radio). April 20, 1926. **Piezo-electric crystal control system.** (Two or more crystals, with synchronizing coils, in parallel).

1,583,417. A. M. Nicolson (Western Electric Co.). May 4, 1926. **Piezo-electric device and method of producing it.** (Rochelle salt crystal attached to loudspeaker).

1,584,490. A. H. Taylor (Wired Radio). May 11, 1926. **Three-phase oscillator.** (Piezo-electric control of three tubes).

1,587,098. H. Whittle (Western Electric Co.). June 1, 1926. **Transformer circuits.** (Tube coupled to piezo-electric loudspeaker).

1,588,176. A. L. R. Ellis (General Electric Co.). June 8, 1926. **Piezo-electric device.** (Methods of avoiding damping due to air in contact with vibrating crystal). Same as German Pat. No. 441,583, and a portion of British Pat. No. 255,463.

1,590,311. A. M. Nicolson (Western Electric Co.). June 29, 1926. **Piezo-electric device and method of producing the same.** (A division of No. 1,438,-965).

1,599,922. T. C. Rathbone (Westinghouse Co.). September 14, 1926. **Balancing machine.** (Use of piezo-electric plate as aid in determining the condition of balance of a rotating body).

1,606,791. J. W. Horton (Western Electric Co.). November 16, 1926. **Oscillation generator.** (Piezo crystal, or other selective system, connected to tube circuits designed for constancy of frequency).

1,608,048. A. H. Taylor (Wired Radio). November 23, 1926. **Piezo-electric crystal control system.** (Relay and protective device to prevent overloading of crystal).

1,608,311. A. L. R. Ellis (General Electric Co.). November 23, 1926. **Oscillator.** (Device to make quartz plate start oscillating easily).

1,609,744. A. M. Trogner. December 7, 1926. **Multiple piezo-electric crystal holder.** (Several crystals on circumference of a drum).

1,617,995. A. L. R. Ellis (General Electric Co.) February 15, 1927. **Piezo-electric device.** (Compensation for changes in frequency caused by varying temperature).

1,619,125. C. W. Hough (Wired Radio). March 1, 1927. **Piezo-electric-crystal apparatus.** (One electrode held in contact with quartz plate by pressure of a column of mercury).

1,619,854. A. Crossley (Wired Radio). March 8, 1927. **Piezo-electric crystal apparatus.** (Crystal plate held in place by guard-ring of insulating material).

1,628,009. A. H. Taylor (Wired Radio). May 10, 1927. **Signal transmission system.** (Potential supply for crystal-controlled generator system, facilitating production of harmonics; also keying system).

1,632,369. A. Crossley (Wired Radio). June 14, 1927. **Radio signaling system.** (Arrangement for control of negative voltage on grid of crystal-controlled transmitter).

1,636,921. A. M. Nicolson (Wired Radio). July 26, 1927. **Piezo audion.** (Crystal plate mounted in evacuated bulb, serving as receiver or emitter of sound).

1,639,817. A. H. Taylor (Wired Radio). August 23, 1927. **Pizo-electric crystal system.** (Calibration of a receiving set by quart resonators, using telephone as indicator).

1,654,184. R. B. Meyer (Wired Radio). December 27, 1927. Frequency-control circuits. (Emergency device to replace piezo-electric crystal in case of failure of latter).

1,654,189. E. L. Powell (Wired Radio). December 27, 1927. Piezo-electric crystal apparatus. (Drum on which are mounted a number of piezo-electric resonators, which can be successively connected to a receiving set for purpose of calibration).

1,654,195. A. H. Taylor and L. C. Young (Wired Radio). December 27, 1927. Modulation system for frequency multiplier circuits. (Crystal-controlled oscillator, frequency-multiplying amplifier, modulation introduced at frequency-multiplying tube).

1,654,196. A. H. Taylor (Wired Wireless). December 22, 1927. Three-phase oscillator. (Piezo-electric control of three tubes, using 1 or 3 crystals).

1,655,974. E. W. Russell and A. F. R. Cotton (Cleveland Trust Co.) January 10, 1928. Piezo-electric device. (Rochelle salt crystal, adapted for use as microphone, receiver, or pickup for phonograph).

British

139,496 M. I. Pupin. May 25, 1921. Improvements in or connected with receivers of high-frequency sound waves. (Two piezo-electric sound receivers so related as to exclude low-frequency disturbances).

145,691. P. Langevin. July 28, 1921. Improvements relating to the emission and reception of submarine waves. (Quartz-metal sender and receiver).

222,150. M. C. Batsel (Metrop.-Vickers Elec. Co. Ltd.). April 9, 1925. Improvements relating to receiving systems for electric signals. (Crystal used as coupling between two tubes).

226,795. P. Langevin and C. L. Florisson. October 22, 1925. Improvements in methods and apparatus for sounding and for locating submarine obstacles by means of ultra-audible waves. (Piezo-electric sender-receiver and recording apparatus).

227,801. P. Langevin. February 4, 1926. Improvements in method and apparatus for the continuous indication or for the recording of depths and distances at sea by means of the ultra-audible echo. (Method involving piezo-electric sender-receiver, for measurement of depth by means of echo).

239,175. W. A. Marrison (Western Electric Co.). September 30, 1926. Improvements in piezo-electric frequency control devices. (Methods of shaping and exciting quartz plates for vibrations of relatively low frequency).

245,810. Western Electric Co., Inc. January 14, 1926. Improvements in telephone transmitters and receivers and the like. (Same as German Pat. No. 430,169, and nearly the same as U. S. Pat. No. 1,574,302).

252,170. E. Giebe and A. Scheibe. Convention date May 15, 1925. Electric tests and measurements. (Luminous effect from quartz bar in evacuated bulb).

252,387. W. H. Whitten (Metropolitan-Vickers Elec. Co., Ltd.). December 2, 1926. Improvements in or relating to scanning devices for television systems. (Mirror vibrated by deformations of piezo-electric crystals).

255,463. A. L. R. Ellis (British Thomson-Houston Co., Ltd.). November 11, 1926. Improvements in piezo-electric devices. (Same as U. S. Pats. Nos. 1,617,995 and 1,588,176; in addition, a device same as German Pat. No. 441,628).

256,611. E. P. Tawil. Published October 6, 1926. Television apparatus; optical systems. (Control of beam of light by piezo-crystal, for television etc.).

258,707. Bell Telephone Laboratories. September 30, 1926. Improvements in radio receiving systems. (Crystal-controlled locally generated oscillations beat with signal waves of a carrier system).

259,174. A. H. Taylor (Wired Radio). May 5, 1927. (Same as U. S. Pat. No. 1,578,296).

260,609. W. G. Cady (Marconi's Wireless Tel. Co., Ltd.). August 11, 1927. Improvements in or relating to means for mounting piezo-electric

crystals. (Methods for centering crystal between electrodes, and for avoiding friction).

261,013. S. Loewe. Convention date Nov. 4, 1925. Thermionic valves. (Crystal-resonator mounted in amplifier or detector tube).

261,040. E. Giebe and A. Scheibe (Radiofrequenz Ges.). Convention date November 7, 1925. Electric measurements. (Low-frequency oscillations and musical note from crystal vibrating in partial vacuum).

261,041. E. Giebe and A. Scheibe (Radiofrequenz Ges.). Convention date November 7, 1925. Wireless signaling. (In receiving system, luminous and acoustic effects from resonator in partial vacuum).

261,042. E. Giebe and A. Scheibe (Radiofrequenz Ges.). Convention date November 7, 1925. Non-contact making relays. (Glow discharge around crystal resonator in partial vacuum utilized as tuned relay).

263,841. Radiofrequenz Ges. and H. Eberhard. Convention date December 29, 1925. Electric measurements. (In parallel with piezo-resonator mounted for luminous effect is a glow-discharge lamp serving as coarse indicator of resonance and to protect the resonator).

264,103. C. W. Rice (British Thomson-Houston Co., Ltd.). January 13, 1927. Improvements in or relating to sound reproducing devices. (Piezo-electric crystal as pickup for phonograph).

264,878. A. W. Hull (British Thomson-Houston Co., Ltd.). June 9. 1927. Method of preparing piezo-electric elements. (Metallic deposit on crystal, subsequently reduced to right thickness.) Same as German Pat. No. 445,046.

266,690. A. W. Hull (British Thomson-Houston Co., Ltd.). March 31, 1927. Improvements in or relating to electric oscillation generators. (Piezo-crystal between filament and screened grid of four-element tube).

268,048. Western Electric Co., Inc. March 28, 1927. Improvements in space discharge tube systems for electric wave signaling. (Crystal in master oscillator circuit, between grid and a point in the anode circuit).

268,367. Telefunken Gesellschaft. October 20, 1927. Improvements in or relating to high-frequency circuits and indicating arrangements therefor. (Air-blast from vibrating crystal closes a contact or moves a lever or mirror).

269,192. Telefunken Gesellschaft. October 20, 1927. Improvements in or relating to piezo-electric devices. (In order to damp vibrations in a crystal or in a body connected thereto, the electrodes of the crystal are connected to a circuit designed to absorb energy).

269,643. W. H. Eccles and W. A. Leyshon. April 19, 1927. Improvements in methods of generating electrical and mechanical oscillations. (Maintaining a body, piezo-electric or other, in vibration, by means of a negative resistance, for example a neon tube).

269,935. E. Giebe and A. Scheibe. Convention date April 24, 1926. Ring-shaped piezo-crystal. (Same as German Pat. No. 450,398).

272,954. Telefunken Gesellschaft. Convention date June 18, 1926. Piezo-electric wave control. (Device to indicate when a crystal resonator is in resonance).

274,660. C. W. Goyder. July 28, 1927. Improvements in or relating to the stabilization of high-frequency oscillations. (Device to keep circuit oscillating if crystal fails).

275,581. J. B. Coleman (Westinghouse Co.). September 8, 1927. Improvements in or relating to high-frequency electric transmitting apparatus. (To send key signals by varying frequency, the airgap of the crystal oscillator is made variable).

276,037. Telefunken Gesellschaft. December 1, 1927. Improvements in or relating to piezo-electric relay and the like devices. (Crystal used as relay or motor employing air-blast effect.)

277,002. H. Eberhard (Radiofrequenz Ges.). Applied for August 31, 1927. (To make vibrations of crystal visible, the latter is mounted close to the wall of a vacuum tube).

277,008. Thomson-Houston Co. October 13, 1927. Thermionic oscillation generator. (Crystal oscillator connected between the grids of two tubes).

Canadian

250,631. A. M. Nicolson (International Western Electric Co.). June 9, 1925. Piezo-electrical transmitter. (Same as U. S. Pat. No. 1,525,823).

French

183,851. J. and P. Curie. May 27, 1887. Electrometer system. (Two piezo-electric bars cemented together so as to give flexure in an electric field).

602,280. Le Materiel Téléphonique. March 16, 1926. Piezo-electric devices. (Method of cutting quartz plate for low-frequency vibrations).

German

430,169. International Western Electric Co. June 12, 1926. (Rochelle salt device for operating loudspeaker. Same as British Pat. No. 245,810).

435,998. International General Electric Co. October 22, 1926. Piezo-electric oscillation generator. (Same as U. S. Pat. No 1,617,995).

441,583. International General Electric Co. March 7, 1927. (Same as U. S. Pat. No. 1,588,176).

441,628. International General Electric Co. March 11, 1927. (Same as part of British Pat. No. 255,463. Composite piezo-oscillator, formed of a mosaic of several quartz plates).

445,046. International General Electric Co. Patent granted June 12, 1926. (Same as British Pat. No. 264,878).

445,052. Telefunken Ges. Patent granted December 13, 1925. Wavemeter and frequency control by means of crystals. (Quartz resonators of progressively increasing lengths, each connected to a small lamp).

450,398. E. Giebe and A. Scheibe. October 7, 1927. Quartz resonator. (Same as British Pat. No. 269,935. Quartz resonator in form of a flat ring with inner and outer electrodes).

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	WASHINGTON	
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Greenleaf Whittier Pickard

Second President of the Institute, 1913

Greenleaf Whittier Pickard

Second President of the Institute, 1913

Greenleaf Whittier Pickard was born in Portland, Maine on February 14, 1877. He is the grandnephew of the poet, John Greenleaf Whittier. Mr. Pickard was educated at Westbrook Seminary, Westbrook, Maine; Lawrence Scientific School of Harvard University; and Massachusetts Institute of Technology.

In 1898–99 he was with the Blue Hill Observatory doing experimental radio work. For several years he was research engineer for the American Wireless Telegraph and Telephone Company. In 1902–06 he was associated with the research department of the American Telephone and Telegraph Company in radio telephony. He practiced consulting engineering in 1906–7. Since 1907 he has been a consulting engineer for and director of the Wireless Specialty Apparatus Company of Boston.

Mr. Pickard is well-known for his early inventions in connection with radio telephony, the crystal detector, loop aerials and direction-finding systems and various static mitigating devices used during the war. He has about one hundred United States and foreign patents.

In 1926 the Institute Medal of Honor was awarded to Mr. Pickard. He has been a prolific contributor to the Proceedings of the Institute. Mr. Pickard is a Fellow of the Institute.

CONTRIBUTORS TO THIS ISSUE

Ballantine, Stuart: (See PROCEEDINGS for April, 1928).

Byrnes, Irving F.: Born at Beacon, N. Y., October 15, 1898. Entered General Electric Test Department in 1918. Engaged in laboratory work on earlier types of tube transmitters, 1919–21. Developed duplex radiophone equipment used on S. S. America for ship-to-shore tests in 1922. Developed crystal control equipment now in use at stations WEAF WGY, KGO and KOA. At present engaged in development work on commercial and military high-frequency transmitters, aircraft radio equipment, crystal control, and train communication apparatus. Associate member of the Institute.

Dellinger, J. H.: Born at Cleveland, Ohio, July 3, 1886. Educated at Western Reserve, 1903–07, George Washington University (A.B. degree 1908); fellow at Princeton 1912–13 from which university received the Ph.D. degree in 1913. Instructor in Physics, Western Reserve 1906–07. U. S. delegate, Interallied Technical Conference on Radio Communication, Paris, 1921; Secretary, U. S. Government Interdepartmental Radio Advisory Committee, 1922–23; Member National Radio Conference, 1922–25; technical secretary, American Section, International Union Scientific Radiotelegraphy. Since 1907 physicist and later chief of radio section of Bureau of Standards, Washington. Vice-President of the Institute, 1924, and President in 1925. Fellow in the Institute.

Friis, H. T.: Educated at Royal Technical College in Copenhagen, 1916; Columbia University, 1919–20. In Research Department, Western Electric Company, 1920–24; Bell Telephone Laboratories, 1925 to date. Mr. Friis' work has been largely in connection with radio reception methods and measurements. He has published papers on vacuum tubes as generators, radio transmission measurements, and static interference. Member of the Institute.

Judson, E. B.: In Naval radio service 1917–1919. Laboratorian at U. S. Naval Research Laboratory, Bureau of Standards, 1919–23. 1923 to date assistant to Dr. L. W. Austin in Laboratory for Special Radio Transmission Research, Bureau of Standards. Junior in Electrical Engineering School, George Washington University. Associate member of the Institute.

Pratt, Haraden: Born July 18, 1891 at San Francisco, Calif. Graduated from College of Mech. and Elec. Engineering, University of California, 1914. Radio operator and installer United Wireless and Marconi Companies, 1910–13. Expert Radio Aide to Navy Department 1915–20. Engineer with Federal Telegraph Company in charge of research, factory and design and construction of chain of radio stations, 1922–27. At present in charge of development of radio aids to air navigation, Bureau of Standards, Washington. Member of the Institute.

Schneider, W. A.: Born at East London, South Africa, January, 1899. Received B.Sc. degree from University of South Africa, 1920; M.S. degree from University of Michigan, 1922, and Ph.D. New York University, 1927. Instructor in Physics, New York University, 1923. Assistant Professor of Physics, New York University 1927——. In charge of electrical measurements and high-frequency laboratory, Washington Square College,

New York University. Member of the Institute and of the American Physical Society and Sigma Xi.

Taylor, A. Hoyt: (See PROCEEDINGS for February, 1928).

Young, L. C.: Born at Danville, Illinois, January 12, 1891. Amateur experimental work 1909–17. Telephone and Telegraph Department, Western Division of Pennsylvania Railroad 1912–17. Experimental radio work under Lt. Cmdr. A. H. Taylor at Great Lakes Naval Station, Ill., Belmar, N.J., and Hampton Roads, Virginia, 1917–1919. Aircraft Radio Section, Bureau of Standards, 1919. Since then research work at Radio Division, Naval Research Laboratory, Anacostia, D.C. Associate member of the Institute.

INSTITUTE ACTIVITIES

April Meeting of the Board of Direction

At the April meeting of the Board of Direction, held in the offices of the Institute on the fourth, the following were present: Alfred N. Goldsmith, President; L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Ralph Bown and Donald McNicol, Junior Past Presidents; Arthur Batcheller, J. H. Dellinger, A. H. Grebe, R. A. Heising, R. H. Marriott, and J. M. Clayton, Secretary.

The following were transferred or elected to the higher grades of membership in the Institute, upon favorable recommendation of the Committee on Admissions: Transferred to the Member grade: F. L. Brittin, C. L. Davis, F. P. Guthrie. Elected to the Member grade: Nathaniel Baldwin, A. L. Loomis, W. A. Schneider, Alexander Senauke, John S. Smith, H. J. Warner.

One hundred and eight Associate members and ten Junior members were elected.

The Institute's Medal of Honor for 1928 was awarded to Professor Jonathan Zenneck of the Techniscke Hochschule, Muenchen, Germany, for his contributions to original research on radio circuit performance and to the scientific and educational contributions to the literature of the pioneer radio art.

Letters from the Federal Radio Commission, requesting certain suggestions from the Institute regarding the allocation of Broadcast channels to Zones and States were read. It was decided that the invitation of the Commission to send representatives to an informal conference of engineers to be held in Washington on April 6th to discuss these matters should be accepted. A Committee composed of the following members of the Institute was appointed: R. H. Marriott, Chairman; J. H. Dellinger, C. W. Horn, and L. E. Whittemore. The joint report of the engineers present at the conference is printed on page 556 of this issue.

An invitation from the Washington Section of the Institute to hold the 1929 Convention of the Institute in Washington, D.C. was read. The Board voted to accept this invitation, and decided that the 1929 Institute Convention will be held in Washington.

Engineering Rise in Radio

The above is the title of a serial account of the rise and progress of radio, from the engineering and invention viewpoints, which has been written by Donald McNicol, past president of the Institute. The story begins in the June, 1928 issue of the magazine *Radio Engineering* published in New York, and will be continued in the monthly issues throughout the year.

Applications for Transfer or Election

It is very helpful if applicants for transfer or election to the higher grades of membership in the Institute (Member and Fellow) will forward a statement of their training and past experience to the persons whose names are used as references in connection with applications. Much time on the part of the Committee on Admissions in communicating with references will be saved.

In general where the Junior or Associate applicant does not know five Associates, Members or Fellows whose names can be used as references, it will be acceptable if these applicants furnish the names of five non-member engineers, or persons who are otherwise engaged in scientific work, who are personally familiar with the applicant and his radio interests.

Institute Meetings

New York Meeting

I. F. Byrnes, of the General Electric Company, presented a paper "Recent Developments in Low Power and Broadcasting Transmitters" at the New York meeting held on April 4th in the Engineering Societies Building, 33 West 39th Street. The paper is printed elsewhere in this issue.

Following the presentation of the paper the following members took part in its discussion: Messrs. F. M. Ryan, Melville Eastham, A. N. Goldsmith, J. H. Dellinger, I. F. Byrnes, E. L. Nelson, and others.

Over three hundred members of the Institute and their guests attended this meeting.

On May 2nd K. S. Van Dyke of Wesleyan University will read a paper "The Piezo-Electric Resonator and its Equivalent Network" before the New York meeting.

On June 6th two papers will be presented before a New York meeting. They are "Developments of Radio Aids to Air Navigation" by J. H. Dellinger and Haraden Pratt, of the Bureau of Standards, and "Aircraft Radio Installations" by Malcolm P. Hanson, of the Radio Division, Naval Research Laboratory.

Atlanta Section

On March 28th a meeting of the Atlanta Section was held in the Chamber of Commerce Building. Mr. Clark of the A. H. Grebe Company presented a paper "Radio-Frequency Receiving Sets."

On April 4th, in the Chamber of Commerce Building, a meeting of the Atlanta Section was addressed by M. O. Mosteller on "Shield Grid Tubes." A second paper by I. F. Byrnes on "Recent Developments in Low Power and Broadcasting Transmitters" was read by C. F. Daugherty.

Buffalo-Niagara Section

A meeting of the Buffalo-Niagara Section was held in Foster Hall, University of Buffalo, on March 14th. William E. Brindley, of the Radio Engineering Department, Westinghouse Electric and Manufacturing Company, presented a paper on "Line Power Operated Radio Sets."

The paper pointed out the public's preference for economies and conveniences of the line-power operated radio broadcast receivers as justification for the further development and production of this type of apparatus.

In order that the relative merits of the two generally accepted methods of accomplishing light-socket operation might be plainly evident, the series filament method was described first. The method, which involves the use of tubes whose filaments are heated by alternating current, was then discussed. This discussion involved the means for lighting the filaments of a-c. tubes, and for obtaining suitable plate and bias voltages; the precautions taken to prevent undesirable couplings between the various amplifier circuits of the receiver; means for locating grid returns at the electrical center of the filaments to keep hum at a minimum; input voltage regulation; line voltage accommodation and utilization of electro-dynamic d-c. excited cone speakers with line-power operated sets.

On April 11th there will be a joint meeting between the Canadian, Rochester, and Buffalo-Niagara Sections in Foster Hall, University of Buffalo.

Canadian Section

The Canadian Section held its March meeting on the fourteenth in the University of Toronto. R. A. Hitchcock, of the Radio Engineering Department, Westinghouse Electric and Manufacturing Company, presented a paper, "Quartz Radio Frequency Standards." Messrs. Soucy, D. De F. Bagley, D. Hepburn, T. R. Rosebrugh, C. A. Lowry, and J. F. Hill discussed the paper.

Seventy-four members attended the meeting.

On Wednesday, May 2nd, Mr. Clark, of the Bell Telephone Company of Montreal, will present a paper on "Carrier Current Communication," in the Electrical Building, University of Toronto.

Chicago Section

A meeting of the Chicago Section was held on March 16th in the Auditorium of the Western Society of Engineers, Monadnock Building. Walter Danley presented a paper "A High-Frequency Filament Lighting Supply System." The paper was discussed by Messrs. Miller, Marco, and Wilcox.

Forty-four members attended this meeting.

Cleveland Section

A meeting of the Cleveland Section was held on April 6th in the Case School of Applied Science. Clayton C. Russell, engineer of station WTAM, presented a paper on "Behind the Microphone of WTAM." The paper described the aerial, station, and equipment of WTAM, the first storage battery operated radio station. Pictures were projected to illustrate fully present and past equipment in use at the station.

Forty-two members attended the meeting.

Detroit Section

The Detroit Section of the Institute held a joint meeting with the Detroit-Ann Arbor section of the American Institute of Electrical Engineers on the evening of March 20th in the auditorium of the Detroit Edison Company. After viewing an interesting motion picture entitled "Voices Across the Sea,"

which portrayed the transatlantic telephone circuit in operation, F. H. Riddle, Chairman, presented the principal speaker of the evening.

Austin Bailey, of the American Telephone and Telegraph Company, presented "The Transatlantic Radio Telephone Circuit." He presented the problem of connecting the respective telephone systems of North America and Europe, separated by over 3,000 miles of ocean. The speaker indicated that the most feasible way of carrying out this connection, for the present at least, appeared to be by the use of radio. Some of the major difficulties involved in making such a connection were pointed out. From the fundamental radio telephone circuit which might be used for communication between two individuals on two sides of the Atlantic ocean, consisting of a radio transmitting station, a radio receiving station, a telephone transmitter, and a telephone receiver for each person, the present transatlantic radio telephone arrangement was developed. A rather detailed account of the wire circuit arrangements, the voice-operated switching device, the single side-band radio transmitter, the receiving wave antenna and the double demodulation radio receiver was given. Lantern slides of schematic circuits and of the necessarily complicated apparatus units were shown and explained.

Los Angeles Section

The Los Angeles Section of the Institute met on March 19th at the Elite Cafe, 633 S. Flower Street, Los Angeles. T. F. McDonough, in the absence of Mr. Wallace, the Section Chairman, presided. R. B. Parrish, Consulting Engineer of Los Angeles, talked on inductive radio interference and power leak hunting. Bert Fox described some interesting short wave experiments.

Twenty-two members attended the meeting.

Philadelphia Section

The Philadelphia Section held a meeting on March 23rd in the Bartol Laboratories, Philadelphia. W. E. Thompson, of the Moore School of Electrical Engineering of the University of Pennsylvania, presented a paper on "Sound." Messrs. Patterson, Wilson, and others discussed the paper.

Sixty members of the Section attended the meeting.

Pittsburgh Section

The first organization meeting of the newly formed Pittsburgh Section was held on April 3rd in the English Room of the Fort Pitt Hotel.

The following officers were elected: Chairman, W. K. Thomas; Vice-Chairman, L. A. Terven; Secretary, A. J. Buzzard, and Treasurer, Anthony Mag.

A paper entitled "Televox" was presented by R. J. Wensley, of the Westinghouse Electric and Manufacturing Company.

Forty-nine members attended this meeting.

Rochester Section

A meeting of the Rochester Section together with the Rochester Engineering Society and affiliated societies was held on March 2nd in the Rochester Chamber of Commerce Building.

Hugh M. Stoller, of the Bell Telephone Laboratories, presented a paper, "Television." Five hundred and seventy-eight persons attended the meeting.

On April 6th a meeting of the Rochester Section was held in the Hotel Sagamore. C. A. Boddie, of the Radio Research Engineering Department of the Westinghouse Electric and Manufacturing Company, presented a paper, "Carrier Current Communication and Control."

Washington Section

The following committees are announced by F. P. Guthrie, Chairman of the Washington Section: Membership Committee: Haraden Pratt, Chairman, G. D. Robinson, C. B. Mirick, Comdr. B. V. McCandlish, Eugene Sibley, J. E. Smith and C. D. Backus. Publicity Committee: R. D. Heinle, Chairman; Carl Butman, Thomas Stevenson, Bert Linz, and S. R. Winters. Meetings and Papers Committee: L. P. Wheeler, Chairman, Captain Guy Hull, and T. Parkinson.

A meeting of the Washington Section was held on April 12th in Picardi's Cafe, 1417 New York Avenue, N. W. Dr. C. B. Jolliffe (Vice-Chairman) presided.

Alfred Crossley, of the Naval Research Laboratory, presented a paper, "Modes of Vibration in Piezo-Electric Crystals." Messrs. Dellinger, Wheeler, Robinson, Jolliffe, Brewington, Hund, Herndon, Heaton, and others participated in the discussion which followed.

Fifty-seven members attended the meeting.

On May 10th two papers will be presented before the Section. The first, by M. P. Hanson of the Naval Research Laboratory, will be "Aircraft Radio Installations." The second, by J. H. Dellinger and Haraden Pratt, will be on "Radio Aids to Air Navigation."

Institute Committees

Radio Advisory Committee, Department of Commerce

Lewis M. Hull has been appointed the Institute's representative to the Radio Advisory Committee to the Department of Commerce, to succeed Professor L. A. Hazeltine.

Subcommittee on Receiving Sets

A meeting of the Subcommittee on Receiving Sets was held at the Institute Office on March 7th. Those present were: J. H. Dellinger (Chairman), E. Austin, E. J. T. Moore, B. Olney, M. C. Batsel, C. A. Wright, J. H. Pressley, G. C. Crom, W. A. Diehl, E. T. Dickey, and V. M. Graham.

The subject of changing the value for Interference Output was brought up, and after some discussion it was decided to change this value under Definition of Terms from 0.000005 to 0.00005 watts (50 microwatts). The report of the Committee on Additional Tests was read and discussed. A number of changes were recommended, and the report was referred back to the Committee to be rewritten in final form. The report of the Committee on Test Procedure was next taken up, and it was decided that some change should be made in the output termination for receiving sets. It was decided to refer this matter to a committee of four, to consist of two members from the Subcommittee on Receiving Sets, and two members from the Subcommittee on Electro-Acoustic Devices.

Dr. Dellinger informed the Committee that he felt it would be impossible for him to continue to serve as Chairman of the Subcommittee due to his having accepted the chairmanship of the Committee on Meetings and Papers.

Mr. E. T. Dickey has been appointed Chairman of the Subcommittee on Receiving Sets.

On April 5th a meeting of the Subcommittee on Receiving Sets was held in the Institute Office. Those present were: E. T.

Dickey (Chairman), R. H. Langley, E. Austin, C. A. Wright, G. C. Crom, V. D. Landon, E. J. T. Moore, and W. A. Diehl.

I. G. Maloff was appointed representative from this Subcommittee to the Subcommittee on Bibliography.

The purpose of this meeting was to coordinate the work which had been done so far along the line of revising the Receiving Set Test Standards. The various sections of the standards were gone over in considerable detail, and recommendations made regarding their improvement. It was decided to put the final writing up of the standards in the hands of the Subcommittee, the membership of which is as follows: E. Austin (Chairman), L. M. Hull, and V. M. Graham. This committee is to review the entire standards for the Receiving Sets Subcommittee, and their report will be circulated among the membership of the Receiving Sets Subcommittee and then presented to the main Committee for further consideration.

A letter was read from the Bureau of Standards in reference to a proposed change in their standard frequency schedules. It was the opinion of the Subcommittee that it would be desirable to rearrange their schedule so that the complete broadcast band of frequencies may be covered in one evening's transmission of standard frequencies.

Subcommittee on Electro-Acoustic Devices

The final meeting of the 1927 Subcommittee on Electro-Acoustic Devices of the Standardization Committee of the Institute was held at Institute headquarters on March 7th. Final action was taken on a number of general terms and definitions that had not been the subject of written discussion. The reports of two of the Subcommittees, one covering definitions of horns and radiators and the other covering suggested symbols to be used in connection with the definitions of Electro-Acoustic Devices, were considered, but not finally acted upon. A report covering all terms and definitions that had been adopted by this Subcommittee, together with a file of unfinished matters, is being prepared for submission to the main Committee.

At the request of the Subcommittee on Receiving Sets the Chairman appointed Melville Eastham and H. A. Frederick as members of a special Subcommittee to cooperate with a similar group from the Receiving Sets Subcommittee with the object

of specifying the nature of the load in the testing of radio receiving sets.

The Chairman expressed his appreciation to the members of the Subcommittee for their cooperation and work in preparing the first set of terms and definitions of electro-acoustic devices to be proposed by an Institute Committee.

Report of Radio Engineers to the Federal Radio Commission

At the request of the Federal Radio Commission, the Board of Direction of the Institute appointed a committee to attend the informal conference of engineers to consider proposed broadcasting channel allocation. The report which follows is a report of the recommendations of engineers present at the Conference, and was prepared by J. H. Dellinger, Acting Chairman of the conference held in Washington on April 6, 1928.

Resolution

It is the opinion of the engineers in attendance that from a radio engineering standpoint, under the provisions of the 1928 law requiring equality between zones, Plan A, submitted for discussion by the Commission, modified as follows, represents the maximum obtainable radio service from the available broadcasting channels in the present state of the art.

	Channels		Full Time Assignments	
	Per Zone	U. S.	Per Zone	U. S.
Class C 5,000 to 50,000 watts	10	50	10	50
Class B 300 to 1,000 watts	18	36	18	90
Class A 0 to 250 watts	4	4	40	200

Discussion

Division into Classes.—The readjustment of station allocations required by the 1928 Radio Law gives the Radio Commission an opportunity to provide the radio listeners of the United States with a grade of radio broadcasting service far superior to that furnished under the present allocation of stations. A redistribution of broadcasting stations among the states will result, if the proposed classification of services be

established, in the satisfactory reception of more programs at a higher signal strength by a greater number of listeners in a larger total area than at present, and will do this with less interference than now exists.

The fundamental change required to bring about any material improvement is to provide a considerable number of channels upon which only one station operates. The reason for this is purely physical fact. Since heterodyne interference extends to many times the distance to which actual program service from a broadcasting station extends, operation of two or more stations on a channel results in an area of destructive interference much greater than the area in which program service is provided. Program service, free from interference, can be furnished at great distances from a station only when the station has exclusive use of its channel.

Since there are only 90 channels available for broadcasting in the U. S., ninety is the upper limit of the possible number of stations giving service at considerable distances.

When two or more stations operate simultaneously on a channel, program service can be furnished at short distances from each station without destructive heterodyne interference within that distance, provided the stations are located at proper distances apart corresponding to the power used. Under these conditions many stations can be operated for short-distance local service on a single channel. Outside the local service areas heterodyne interference will prevent satisfactory reception.

Sections of the country remote from centers of population can not be given service except by the stations first mentioned, which have exclusive use of their channels (Class C).

It follows that the country as a whole can be given the service it demands only by having more than one class of station, (1) long-distance stations operating on exclusive channels, (2) shorter-distance stations, operating on shared channels. Considering the broadcasting needs and development in this country, it is apparent that the second class can advantageously be subdivided into stations of moderate distance range (Class B) and small stations of very small distance range (Class A).

Number of Channels in Each Class.—The number of channels (50) indicated for Class C stations is the minimum that should be provided, in view of the far greater service, both distant and local, that will be rendered by such channels, owing to

to the absence of heterodyne interference and the consequent possibility of the use of greater power. The distribution of the remaining 40 channels between Classes B and A represents the best judgment of the engineers from present information. A further study should be made of this point on the basis of service requirements of various areas of the country. It is believed that the final answer on this point will not depart widely from the figures given.

Duplication of Assignments Per Channel.—It is clear that the stations depended upon for service over large areas must that the stations depended upon for service over large areas must operate on heterodyne-free channels and that therefore there must be only one assignment to each Class C channel.

The moderate-distance (Class B) and short-distance (Class A) channels may each be used by a number of stations in simultaneous operation, since the only desideratum is good service within the local service range of each station. The power required for moderate-distance service (Class B) will not permit as much duplication of stations on one channel as will the smaller power required for short-distance service (Class A).

The amount of duplication recommended is: for each Class B channel, on the average of two and a half assignments in the U. S. (i.e., the assignment of every other channel in each zone); and for each Class C channel, 50 assignments in the U. S. (10 in each zone).

The limitation to two and a half assignments for each Class B channel is determined by the geographical circumstances of the two smallest zones (1 and 2), together with the requirement of the law of equality between zones. Points in zones 1 and 2 average less than 500 miles apart, a distance too small to permit the assignment of any one channel in both zones with recommended power.

Equality with Respect to Classes.—The provisions of the law requiring equal distribution among zones, and, according to population, among the states, of station licenses, frequencies, time, and power, must be applied separately to each of the three classes of stations mentioned. This results from the inclusion of the number of licenses as one of the elements of equal distribution.

Station Power.—In order to merit the use of a Class C channel a station must be competent to serve a large area. It follows that no Class C station should be allowed to operate with

less than 5,000 watts power. The only upper limit for this class need be that fixed by the production of inter-channel interference, and, in consideration of the geographical distribution possible, may be 50,000 watts at the present time.

For the moderate-distance (Class B) channels powers of 300 to 1,000 watts will give satisfactory service, and for the short-distance (Class A) channels power should not exceed 250 watts per station because of the extensive duplication permitted.

As an exception to these general recommendations for Classes B and A, it is noted that where two or more stations operating on the same channel are all increased in power by the same factor, their heterodyne-free service ranges will be substantially unaffected and a better signal (with respect to noise interference) will be delivered within each service area. This will be at the expense of producing a stronger heterodyne whistle outside the service areas of the two stations concerned.

Time Division.—The expedient of time division does not in general lead to superior service to the listener. It is inherently uneconomic. Where several stations in an area are now dividing time, the duplication of plant and overhead necessarily results in poorer service than would result were these stations to be consolidated into a single station using all the time.

For the Class C stations praticularly, time division should not be allowed. An exclusive (Class C) channel is capable of delivering such excellent service over large areas that care should be taken to restrict the possible service from these channels by an uneconomic arrangement such as time division.

For the Class B and Class A channels there will doubtless be local conditions demanding, and perhaps justifying, time division in spite of its inherently uneconomic nature. However, the application of time division has been made difficult under the terms of the new law. Since the law requires equality of the number of hours and licenses among zones, and, according to population, among the states within each zone, if time is divided on a given channel among several stations in any one state, this division must be duplicated on some channel in every other zone and proportionally in every state.

The same difficulty will exist in any attempt to divide time between stations located in different zones, as might be sought, e.g., to take advantage of the time difference between the east and the west coasts. Time division between stations in widely separated localities is subject to the further objection of seriously complicating the maintenance of the proper frequency separation between stations in each of the localities to minimize inter-channel interference.

OBITUARY

With deep regret the Institute announces the death of

Thomas R. Bristol

Mr. Bristol served both as President and Secretary of the Atlanta Association of Radio Engineers, and the growth of that organization which permitted its incorporation into a Section of the Institute of Radio Engineers was mainly due to his untiring efforts.

As an official of the Georgia Power Company, a leader in the activities of the Atlanta Section, and as a friend, he won the respect and admiration of all with whom he came in contact. The Section and his many friends throughout the South feel the loss of this ardent worker most keenly.

STUDIES OF HIGH-FREQUENCY RADIO WAVE PROPAGATION*

By

A. Hoyt Taylor and L. C. Young

(Naval Research Laboratory, Bellevue, Anacostia, D. C.)

Summary—*Studies of multiple signals of high frequency have been made upon a quantitative basis with reference not only to the round-the-world signals (sometimes called echo signals) but to nearby echoes which have a very much shorter time of arrival. A method of predicting in advance the the likelihood of round-the-world echoes occurring between any two different stations has been worked out.*

Further studies of nearby echoes show a remarkable retardation on very high-frequency signals coming from the Rocky Point stations to Washington, these signals traveling an actual distance varying from 2900 kms. to over 10,000 kms., although the great circle distance is only 420 kms. The apparent violation of the skip distance law by these stations as observed in Washington has been explained. The nearby echo signals have been tentatively assumed to be due either to reflections from a heavily ionized region in the neighborhood of the magnetic poles or more likely to be due to scattered reflections thrown backwards from the first and second zones of reception, which follow the skip distance region. This throwing back of the signal thus permits under certain conditions the reception of the signal on very high frequency within what is really the skipped zone, the signal having entered this zone by a very indirect route and with a considerable time retardation as compared with the direct route.

Influence of both nearby and round-the-world echo signals upon various types of radio communication have been briefly discussed.

FOR some time the Naval Research Laboratory has been interested in the study of so-called echo signals, particularly those which travel entirely around the world. Early in the fall of 1927 it was decided to put these studies on a quantitative measurement basis in the hope of obtaining information which would be of value in the more general study of wave propagation in theory and practice. It was expected that these studies would throw additional light upon the structure of the Heaviside layer and its daily and seasonal variations. Early in these studies, however, certain phenomena manifested themselves which gave added incentive.

There are, of course, many questions bearing on wave propagation that are by no means cleared up at the present time but among them are several which are of considerable theoretical

* Original Manuscript Received by the Institute, March 12, 1928.
* Presented before meeting of Washington Section of the Institute, March 8, 1928.

and practical interest and which may be mentioned at the outset of this report:

(1) Why are signals usually received during the daylight hours in Washington from the Radio Corporation's group of stations at Rocky Point, operating on frequencies so high that Washington is well inside of the skip distance zone? This point was raised in a paper by Dr. A. H. Taylor[1].

(2) Why do distant stations show what may be termed a normal time interval for round-the-world signals, whereas the Radio Corporation group as measured in Washington invariably show time intervals notably shorter than the normal time interval? These two questions are intimately related.

(3) Can any further explanation be given of the fact that high-frequency signals usually show the worst fading at moderate distances and much less fading at the extreme limits of the range?

(4) What influence will echo signals, so-called, or multiple signals, which is a better designation, have practically upon low-speed telegraphy, high-speed telegraphy, telephony, facsimile transmission and television?

(5) To what degree can directive transmitting and receiving systems be expected to suppress multiple signals?

(6) During what hours of the day and what seasons of the year are different types of multiple signals to be expected?

(7) Finally, whether there is anything in the nature of echo signals which is in contradiction with the general theoretical principles already laid down.

This paper being only in the nature of a preliminary report we do not by any means pretend to give final answers to all these questions, but it is believed that it will throw considerable light upon some of them and introduce possibly some new speculations which may at least stimulate constructive criticism, comment, and further work along similar lines in other parts of the world, which we believe to be the only means of finally settling some of these questions.

Turning to the consideration of multiple signals which are due to waves which have traveled around the world in the sense opposite to the direction from transmitter to receiver, a number of observations will be presented giving the equivalent time for the signals to pass entirely around the world, these times

[1] Proc. I.R.E., 14, p. 539, August, 1926.

being based on the observed time differences between the direct and the rearward signal; namely, between the signal coming by the short route transmitter to receiver and the one coming by the long route in the opposite way around the world.

It should perhaps be explained at this point that records were made upon a Westinghouse multiple tower oscillograph and by the use of two superheterodyne receivers using different transfer frequencies and usually operating on antennas of different directivity. Another way to handle these two receivers is to use the same transfer frequency and the same heterodyne, thus avoiding whistling between the receivers. The fact that the higher frequencies from San Francisco failed to go around

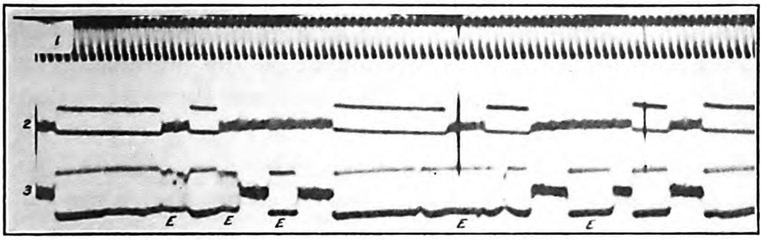

Fig. 1—Around-the-World Signals from PQW, Lisbon, 9:15 A.M., December 8, 1927. Frequency 19,180 kc.

seems to agree with the theoretical deduction that the height of the layer seldom averages low enough on the necessary great circle for 16,700 kc. and therefore would rarely, if ever, be low enough to carry around the still higher frequency.

The expression "height of the Heaviside layer" as used in this paper means the height of the equivalent reflecting layer. It does not mean the height at which the electron concentration is the maximum. In reducing path retardations to a time basis the velocity of light in free space has been used so as to put all observations upon simple comparative basis. No allowance has been made for the number of reflections which a wave might make in going from one point to another either by direct or indirect routes. Such corrections would slightly modify the time figures but not by any large amount.

Fig. 1 shows an observation taken on PQW, Lisbon, the multiple signal being marked E. The frequency here under observation is approximately 19,000 kc. The time of this particular observation was 9:15 A.M. Washington time. This figure

shows multiple signals which are almost of the same order of intensity as the direct signals and almost equally clearcut. The observation was taken also in such a way as to indicate the direction from which the echo signal came. The line 1 on the top of the figure is the timing line from a 100-cycle tuning fork. Line 2 which shows the direct signals practically without echoes was taken on an antenna highly directive towards Lisbon, whereas line 3 which shows the echoes as well as the direct signals was taken on an antenna having strong westerly directivity. This directivity is, of course, not sharp enough to shut out the direct signal from Lisbon but is good enough to emphasize greatly the strength of the echo signal. Our general conclusion from a great many studies of this sort is that echo signals from European stations generally come in during the fall and early winter months best during the middle of the morning hours, say from 0900 to 1200 zone five time[2] and from the southwesterly

TABLE I

ECHOES ON DISTANT STATIONS

Call Letter	Location	Time of Observation	Observed Time	Around World Equivalent	Freq. kc. (approximately)
NPG	San Francisco	1310	0.1120	0.1393	16700
"	U. S. A.	1410	0.1130	0.1405	16700
"	"	1410	0.1130	0.1405	16700
"	"	1420	0.1125	0.1400	16700
"	"	1420	0.1120	0.1393	16700
"	"	1420	0.1125	0.1400	16700
AGA	Nauen	0905	0.0920	0.1382	20250
"	Germany	0908	0.0920	0.1382	20250
"	"	0915	0.0920	0.1382	20250
"	"	0915	0.0910	0.1367	20250
PQW	Lisbon Portugal	0915	0.0980	0.1374	19180
"	"	0915	0.0990	0.1386	19180
"	"	0915	0.0975	0.1352	19180
"	"	0920	0.0975	0.1352	19180
"	"	1110	0.0985	0.1380	19180
"	"	1110	0.0980	0.1374	19180
"	"	1110	0.0985	0.1380	19180
"	"	1010	0.0985	0.1380	19180
PCPP	Kootwijk Holland	0920	0.0950	0.1362	18600
"	"	0930	0.0925	0.1326	18600
"	"	0931	0.0940	0.1347	18600
"	"	0930	0.0940	0.1347	18600
"	"	0905	0.0960	0.1376	18600
"	"	0905	0.0960	0.1376	18600
"	"	0905	0.0970	0.1390	18600
"	"	0912	0.0965	0.1383	18600
"	"	0855	0.0930	0.1333	18600
"	"	0855	0.0940	0.1347	18600
"	"	0858	0.0965	0.1383	18600
PCRR	Kootwijk Holland	0926	0.0960	0.1376	14600
"	"	0928	0.0960	0.1376	14600
"	"	0928	0.0940	0.1347	14600
"	"	0928	0.0940	0.1347	14600
"	"	0940	0.0950	0.1361	14600
"	"	0940	0.0960	0.1376	14600
"	"	0940	0.0940	0.1347	14600

[2] (Editor's note: Zone five time is 75th meridian time—Eastern Standard Time.)

direction. The same is generally true of the Rocky Point group of the Radio Corporation.

On the other hand, echoes from the Naval Station NPG at San Francisco on a frequency of 16,700 kc. rarely are picked up on a collector with strong westerly directivity and when they are so picked up they indicate by their timing that they have traveled around the world in the same direction as the direct signal. Such echoes, however, are usually weaker than the reverse direction type but can be at times extremely annoying although those times are more limited in duration and do not occur as often during the year. Directivity is no defense whatever against echo signals of such a nature but it is a great defense against echoes coming in the reverse direction, as Fig. 1 plainly indicates. Table I gives a summary of measurements on various distant stations; the first column giving the call letters of the station; the second, its location; the third, time of observation; the fourth, the actual observed time interval between direct and echo signals; and the fifth, this same time interval reduced to the equivalent interval for a transit entirely around the world.

A study of this table shows variations in the time of round-the-world transit which are clearly outside of the errors of observation, but nevertheless the figures are fairly consistent and with few exceptions can be correlated with a reasonable Heaviside layer height. It is not believed that these round-the-world signals follow a curvilinear path around the world but that they encounter the layer at intervals and are turned down and reflected back up again from the earth's surface. Variations in layer height will therefore introduce small variations in the path's difference between the direct signal and the echo signal, and indeed we regard the failure of these values to show greater constancy as clear evidence of this manner of progression of the waves. From an average of round-the-world time values a figure can be set which can be called the normal time interval for round-the-world signals in the frequency band between 16,000 and 22,000 kc. Without an intimate knowledge of the exact number of jumps which the wave makes both for direct signal and echo signal, it will be impossible to say exactly what this value should be for a given height of the layer as there are a number of possible combinations. Further than this we cannot go although it is likely that there is some variation in this time interval with the frequency.

Turning for a moment to the question of the season of the year and time of day when signals of this nature should occur, it is possible to make certain theoretical predictions which, so far as our study has progressed, have been verified by practice. There is considerable evidence in favor of the assumption that the principal signal energy comes along a great circle path both for the direct and for the round-the-world signals. This is a very difficult matter to subject to experimental verification, nevertheless in the absence of information to the contrary we shall assume for the present at least that the main signal energy does so come to the receiver, our main reason in making such assumption being the action of collector systems of the Beverage type with a fair degree of directivity. This being so, it can be seen that a round-the-world signal originating in the northern hemisphere during winter time must, unless a great deal of the preceding theoretical work be wrong, get through the northern hemisphere during the light hours and do its dark hour transit in the southern hemisphere where it is summer. Otherwise, radiations starting out along great circles which would violate this simple principle would surely encounter regions where the Heaviside layer was too high for such frequencies to come down again to earth.

So far as directional observations have been taken on echo signals this simple principle has been substantiated and if one takes the globe and traces upon it the possible paths of great circles which would bring the echo signal back to the near neighborhood of the transmitter, it is quite plain that signals originating in the northern hemisphere and observed regularly close to the transmitter will show echo signals only during certain hours of the day and that these hours are by preference between 0800 and 1200 local time in the winter. It is also clear, if one lays off the possible great circles which would meet the simple condition of transversing their night path in the summer hemisphere, that the seasonal change from winter to summer will gradually push this echo period towards the afternoon, and indeed this is generally true of the observations today. We venture to say that in the mid-summer period (which we have not yet had opportunity of carefully observing) the reverse will exist and the favorite echo period will be between 1900 and midnight for the same stations.

Now, when transmitter and receiver are widely separated more complicated conditions govern the situation, but these conditions also are subject to analysis. For instance, if one considers the situation between San Francisco and Washington and remembers that here we are practically limited to one great circle and not to a choice of any one that may be possible for the signal to find its way around on, it can be seen that there are only a few periods of the day and of the year when the great circle San Francisco-Washington traced backwards could be expected to pass through a region with a sufficiently low Heaviside layer height. One of these periods is during the fall and early winter months at 1500 Washington time, and it is significant indeed that very few echoes have ever been found on NPG at any other time of the day and, so far, of the year. It is believed also that the occasions on which, even at these times, the layer is low enough to give this effect without losing the reverse signal at some point of the route, are relatively few and far between, which is also borne out by observations. Further evidence on these points is that the Radio Corporation stations on frequencies considerably higher and located not far from NPG have failed to give any echoes at all except at extremely rare and transient intervals.

Another important observation can be made here upon the performance of stations to the south of us. Strong signals are received here from the Radio Corporation station at Bogota, HJG, both in the 13,700 kc. and in the 27,400 kc. bands at certain hours of the day, but never have echo signals been observed on this station or on any of the stations in the Argentine and Brazil on similar frequencies. Theoretically, no echoes would be expected because such echoes would have to pass over both poles and one or the other is bound to be in total darkness during its winter period even in the daytime, and therefore probably has too high a layer to permit such frequencies to be carried around. Whether it will be possible during the equinoctial period to observe echoes on such southern stations is a point of considerable interest but upon which we have so far no positive information. It is barely possible that during these periods the average layer height might be low enough to get certain frequencies around. It is also quite clear that only very high frequencies, namely, above 14,000 kc., are suitable for sending frequencies entirely around the world, except

at certain very limited periods, because the signal must traverse thousands of miles of daylight and only the very high frequencies have the low absorption necessary to accomplish this.

Fig. 2 shows a record on CRHB, Cape Verde Islands, on approximately 16,500 kc. Unfortunately the timing line was

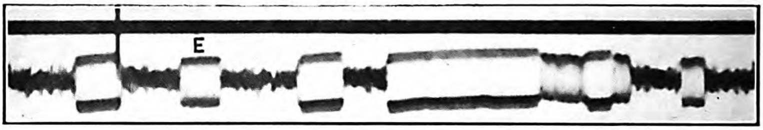

Fig. 2—Around-the-World Signals from Cape Verde Islands, 1 P.M., January 3, 1928.

omitted in this photograph accidentally so the measurement is not available, but the record shows a clear enough echo which is evidently of a reverse direction signal. The more southerly position of the Cape Verde Islands shifted the favorite echo period from the morning hours toward afternoon, this record being taken at 1300 zone five time and strong echoes being observed for at least one and a half hours later. All this in-

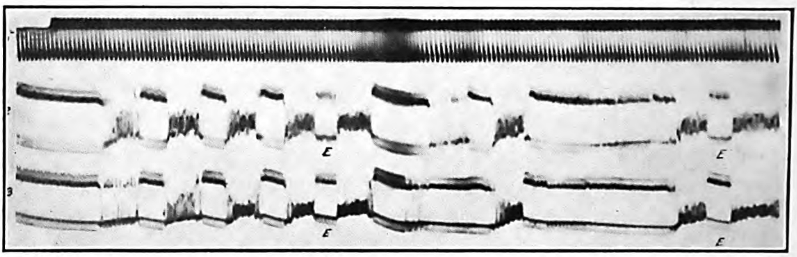

Fig. 3—Around-the-World Signals from 2XBC, 11:15 A.M., December 2, 1927. Sending ABC.

formation checks in with the general idea that echo signals must make their night transit in the summer zone and is in agreement with the general theory that the summer night layer height is notably lower than the winter night layer height except in the tropics where there is not so much difference between summer and winter. Again on this point the Naval Stations in the tropics have noted that they could continue on into the dark hours with very high frequencies over certain

long distance circuits—much later into the night than can be done in the northern hemisphere except in the northern summer.

If this simple rule then is noted, it is not difficult to predict at what time of day echo signals may be expected and how that time of day varies with the time of year, and it is quite clear that when stations are geographically widely separated the times of day and times of year will be very much more limited than when the observations are taken near the transmitter. It is also clear that between certain stations there is little likelihood of echoes of this type ever being observed. Taking the figure 0.139 as the general average for round-the-world signals in the frequency band from 16,000 to 21,000 kc., a marked deviation is at once observed when we consider the observations on the Radio Corporation group.

Fig. 3 shows a typical record out of a large number taken on 2XBC at Rocky Point on a frequency of approximately 20,000 kc. The timing line 1 is made by a 100-cycle fork. Line 2 shows the signal received on a collector of directivity to the northeast, the signals marked *E* being those echo signals that are plainly discernible and identifiable. Line 3 shows the echoes

TABLE II

Around the World Signals—Rocky Point Group

2XBC Time of Observation	Approx. 19,900 kc. Echo Time	WLL Time of Observation	Approx. 18,000 kc. Echo Time
1010	0.1290	1035	0.1105
1145	0.1250	1035	0.1050
1145	0.1240	1035	0.1080
1145	0.1260	1035	0.1100
1225	0.1265	1145	0.1265
1225	0.1250	1154	0.1265
1250	0.1260	1155	0.1270
1255	0.1260	1230	0.1275
1255	0.1250	1235	0.1280
1255	0.1250	1248	0.1280
		1248	0.1285
		1147	0.1260
WTT	Approx. 18,500 kc.	WBU	Approx. 21,200 kc.
1115	0.1250	1025	0.1265
1120	0.1250	1037	0.1275
1034	0.1270	1055	0.1205
1050	0.1250	1012	0.1195
1320	0.1200	1130	0.1100
1005	0.1265	1025	0.1250
1040	0.1250	1125	0.1260
1048	0.1265	1000	0.1260
1120	0.1260	1240	0.1260
1025	0.1260		
WIK	Approx. 13,800 kc.		
1200	0.1230		
0920	0.1270		
0920	0.1280		
0920	0.1200		
1230	0.1110		
0937	0.1330		
0945	0.1260		

plainly discernible on a collector of pronounced westerly directivity. The fact that the echoes are stronger and more clearcut on the westerly collector indicates that the echoes are coming from a general westerly or southwesterly direction. The timing interval instead of being 0.139 seconds is 0.124 seconds.

Table II is a summary of observations taken on the Rocky Point group. An inspection of this table shows that in no case do the signals appear to take the normal time interval to go around the world. Moreover it is seen that the observed time intervals vary from 0.110 to 0.129. Not only are these time intervals decidedly too short, but they show a greater variation than several observations on distant stations. These observations are indeed of great interest. Our first thought was to explain these observations by assuming that the round-the-world signals

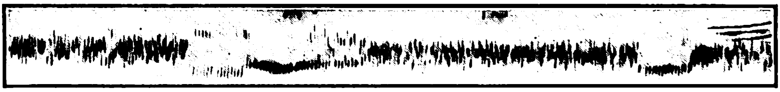

Fig. 4—Around-the-World Signal from NKF.

followed some course differing from the great circle route by an amount corresponding to the time interval, but we were unable to think of any physical mechanism which could produce such a path. It would be much easier to think of conditions which would make the time interval too long than to think of conditions which would make it too short. It had long been suspected that the so-called direct signals coming from the Rocky Point group on these frequencies and which apparently violated the skip distance law were not really direct signals. Means were therefore investigated to throw further light upon this matter.

This Laboratory placed one of its own high-power transmitters on 18,500 kc. and obtained the consent of the Carnegie Institution to take observations at their laboratory on the edge of Rock Creek Park some ten miles distant. Fig. 4 shows a typical echo signal on our own station NKF and with the time interval 0.1390 second. It should be explained that since we had control of the transmitter in this case we were able to arrange it to send extremely short dots of the order of duration of about one-hundredth of a second and widely spaced. This was accomplished by punching a transmitter tape with the letter *E* with ten or fifteen spaces between each letter and then running

the tape through a Creed transmitter at high speed. Roughly, these dots came about three to the second.

Table III shows the results of round-the-world signals from NKF taken at the Carnegie Institution ten miles away, meas-

TABLE III

AROUND THE WORLD SIGNALS FROM NKF.

	Front of Dot	Rear of Dot
	0.1401	0.1382
	0.1380	0.1382
	0.1396	0.1375
	0.1406	0.1385
	0.1387	0.1376
	0.1386	0.1365
	0.1397	0.1400
Average	0.1393	0.1387

Average—front and rear, 0.139 Sec.

urements being given both from the front end of the dot and rear end of the dot which show slight differences which seem to give slightly higher values for the front of the dot than the rear of the dot. We are not willing to lay any special weight

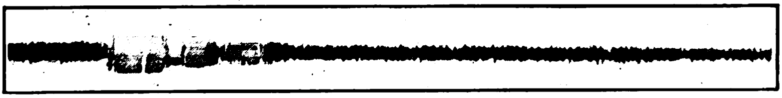

Fig. 5—Reflections from NKF.

upon this timing difference as it is quite small, but at any rate the observations show remarkably little variation and give a general average close to the general average or normal value for distant stations, differing thereby markedly from observations on the Rocky Point group. It was therefore concluded that the so-called direct signals from Rocky Point which appear to violate the skip-distance law, being received well through-

TABLE IV

SUMMARY OF REFLECTIONS FROM NKF

0.0111	0.0207	0.0311
0.0116	0.0212	0.0312
0.0118	0.0222	0.0330
0.0119	0.0224	0.0335
0.0131	0.0226	0.0335
0.0138	0.0232	0.0335
0.0142	0.0232	0.0343
0.0149	0.0232	0.0345
0.0163	0.0233	0.0349
	0.0239	0.0355
0.0181	0.0241	0.0358
0.0184	0.0243	0.0358
0.0185	0.0243	0.0359
0.0188	0.0296	0.0360
0.0190	0.0258	0.0360

out the daylight hours but fading out in the very late afternoon and coming in again just before dawn, were not coming direct at all but were themselves echoes or multiple signals of an entirely different character from round-the-world signals. But if this were true, our own station should also show additional multiple signals or echoes very close to the reference signal or ground wave. Such indeed proved to be the case, as is shown by Fig. 5 where very short time echoes are clearly shown of the values 0.0138, 0.0235, and 0.0345 respectively.

Table IV is a summary of these nearby echoes found on NKF. We therefore concluded that the so-called direct signals from Rocky Point which appear to be violating the skip-distance law were similar echoes, but we sought for definite means of proving this point and finally hit upon means for doing so. The Rocky Point transmitters like most high-frequency transmitters having frequency multipliers, for instance, one of them transmitting on 18,000 kc. has somewhere in the preceding stages considerable energy on 9,000 kc. which is weakly coupled to the antenna system but which should be capable of giving recordable signals in Washington. Examination of this matter shows that such was indeed the case and observations were taken on 2XBC and WLL on this basis and on the following assumptions; first, the lower or half frequency could not be an echo signal because it would have to travel too far for such a frequency to give a good signal in broad daylight, our shortest echoes of this character indicating a path retardation of the order of 2,900 km. Considering the weak radiation on relatively small power put out by the transmitter on this half frequency it is hardly possible that so perfect a signal could have been received at Washington by any other than a normal sky wave route, which means for a Heaviside layer height of under 200 km., an approximate distance from New York to Washington for this signal of about 500 km. Of course, the same relay keyed both the half frequency and the high frequency and no question of relay lag was herein involved.

Fig. 6 shows a typical record on 2XBC clearly showing that the high frequency not only arrives later by 0.007 or 0.008 of a second but that the signals are strung out somewhat longer than those arriving on the lower or half frequency. The two frequencies in question were approximately 18,900 and 9,450 kc. In connection with the observations on NKF this con-

clusively establishes the fact that these high-frequency signals from the Rocky Point area are not violating the skip-distance law at all. Nor are they ground waves or scattered waves which could not possibly be accounted for on this timing basis, but

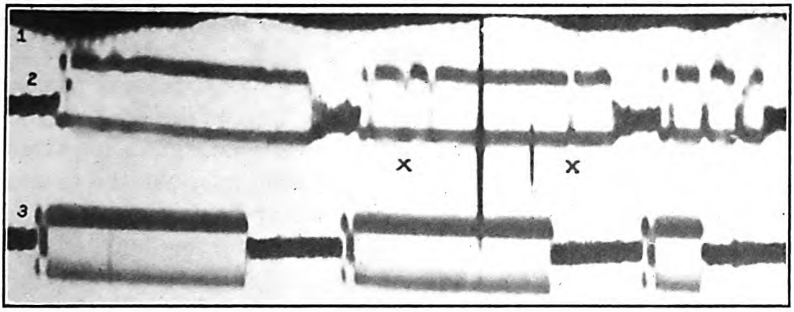

Fig. 6—2XBC, 12:50, February 29, 1928.

they are signals thrown into the skip distance area from a point or a number of points well outside of that area. If we suppose that there are several such echoes of which we have sometimes identified as many as three on NKF, it will be seen that the first echo with the shortest time interval will be the one that starts the recorded dot in the record of the higher frequency. If, therefore, we measure from the beginning of the half frequency dot to the beginning of the high frequency dot we get a time interval corresponding to the nearest echo or the shortest reflection.

Table V summarizes the values obtained on this shortest reflection or echo, which has a value between 0.007 and 0.008

TABLE V
Summary of Shortest Reflections 2XBC

0.00834	0.00866
0.00857	0.00805
0.00802	0.00853
0.00882	0.00808
0.00835	0.00700
0.00846	0.00738
0.00803	0.00750
0.00888	0.00724
0.00823	0.00776
0.00814	0.00765
0.00833	0.00755

of a second and corresponds to a signal which has traveled some 2400 km. further in coming to Washington than the half-frequency signal has. Since the half-frequency signal itself has traveled approximately 500 km. the total distance traveled

by the high-frequency signal in coming from Rocky Point to Washington is about 2900 km. It is very suggestive to know that the average daylight skip distance on these frequencies is about half of this figure. This point will be considered in detail later on. There are plainly to be observed in some of these photographs on the Rocky Point stations interference patterns produced by subsequent echoes.

On Fig. 6 at the point marked X such an interference pattern is shown. Obviously, it will be difficult from the observations made thus on regular traffic and not on special signals to extract the values for intermediate reflection intervals, but the longest interval may be determined by measuring from the end of the half-frequency dot to the end of the corresponding high-frequency record. There is also the possibility of occasional round-the-world signals creating some confusion in these latter type of measurements. For instance, in Fig. 7 the point marked R

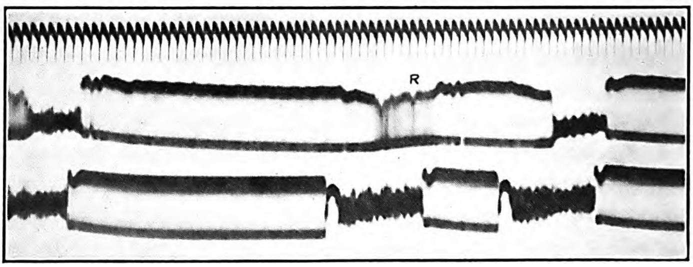

Fig. 7—Around-the-World Signal from 2XBC, 9:10 A.M. February 29, 1928.

is clearly not a nearby, but a round-the-world signal. It is preferable to take these records on nearby echoes at such times when round-the-world signals are not occurring. Apparently then, it is definitely determined that these high-frequency signals within the skip-distance band, New York to Washington, are not violating the skip-distance law but as can be seen from Table V on the nearest reflection and Table VI which gives the list of other and somewhat longer reflections, that we have yet to explain where these reflections come from, both for Rocky Point stations and for NKF. We do not propose to attempt to answer this question at the present time, but will only make a few suggestions as to possible explanations, leaving the final decision to the future when other studies with special

reference to directivity and at other localities widely separated from our own may throw further light upon these points.

It may appropriately be mentioned at this time that since electrons diffuse much more readily along the lines of magnetic force than they do athwart them, there should be considerably more electronic concentration at the neighborhood of the north magnetic pole and that the distance to the north magnetic pole is about right to account for the intermediate set of echoes. The nearby set of echoes might be thought of as being due to reflections of the wave from rough portions of the earth's surface directly back on itself, so to speak, and coming from the first zone of reception; namely, from just above the edge of the first skipped distance. If we look to the southward of Rocky Point on the map we find a rough, more or less mountainous region at the correct distance to be in a position to throw such signals back into the skipped zone between Washington and New York. Equally well, if we look to the northward in Labrador we can find such a zone. It is also true that another group of reflections have a time interval suitable to be echoes from throwback signals from the rough country in the Rocky Mountain area. But the last set of reflections corresponding to time intervals of the order of 0.035 of a second and to a distance of the order of 10,000 km., is difficult indeed to account for. One would have to go south to the Andes or northeast to Switzerland in order to get a similar region adapted for throwing back such signals. We confess, however, that at the present time we are not aware of any reasonable explanation of these nearby echoes which we have found on the Radio Corporation's group and, to the same order of magnitude, upon our own station. We confess also that we have put forth the polar electron concentration idea and the backward reflection idea with no extreme faith in its probability. At the same time we believe that it is not at all impossible for these things to occur. We are naturally

TABLE VI

Long Reflections on 2XBC

0.0152	0.0214	0.0321
0.0165	0.0282	0.0311
0.0178	0.0264	0.0338
0.0141	0.0255	0.0300
0.0169	0.0254	0.0345
0.0197	0.0278	0.0325
0.0190	0.0264	0.0314
0.0126	0.0262	
0.0106	0.0226	
0.0137		
0.0173		

reluctant to entertain the idea of an additonal layer with extremely high electron concentration and of so great a height, namely 1400 km., which would be necessary to explain the path retardations on these nearby echoes by reflections and multiple reflections from such a layer. The existence of such a layer besides being utterly unaccountable on any known physical basis will be too much in contradiction with well-known and well-established facts concerning the behavior of radio waves in various parts of the spectrum. If we are forced to accept the backward reflection idea it can be seen that the nature of these reflections will frequently be such as to give the effect of bad scattering and no definite direction to such signals. Under special geographical conditions and at very high frequencies where long skip distance is obtained before the main wave can find anything that can throw it back upon itself, a certain degree of directivity on certain signals might be observed.

Turning now to the seven questions outlined in the first part of this report, it is perhaps worthwhile to see what answers or partial answers can be given to them in the light of what we have so far determined as to the nature and timing of both nearby and distant echo signals. (1) It is clear that the signals from the Rocky Point group of stations as received in Washington on very high frequencies do not violate the skip distance law nor are they signals formed by a scattering in the normal Heaviside layer. The timing intervals indicate that they come from distances between 2500 and 10,000 km. in making the transit from Rocky Point to Washington, although the straight line distance is only 420 km.

(2) The abnormal time interval on the Radio Corporation group at Rocky Point is fully explained and brought into unison with observations on our own station and in fair agreement with those of distant stations, as soon as we recognize that the signal which arrives first is itself an echo signal of variable time interval corresponding to the above mentioned path retardations.

(3) Violently fading of high-frequency signals at points only moderately distant from the transmitter may be partly due to nearby echo signals which overlap and in continually shifting phase thus contribute to this fading. At points of very much greater distances the nearby echoes would be much reduced in strength and of corresponding lesser importance in producing fading effects.

(4) All of these echo signals can produce under suitable conditions disastrous effects upon almost all known forms of radio communication. The long time echoes coming as they do around the world can usually be taken care of by directive receiving systems, but if short time echoes occur it seems likely that they will come from various directions and directive receiving systems will not be of so much advantage. Upon low speed telegraphy the short time echoes will have little or no effect, their main results being to produce a slight fuzziness of the signal. But upon telephony they will have a very disastrous effect and probably equally annoying effects for facsimile transmission and television. Fortunately, we shall usually be mainly interested in telephony, facsimile transmission, and television only over long distances. The evidence herein presented is perhaps one more argument for the reservation of such frequencies for long distance work only.

(5) Directive receiving systems can obviously be used to great advantage with certain types of echoes but can hope to do but little with others. Fortunately, in long distance work we may expect comparatively little trouble from nearby echoes on account of their lesser intensity. Directive transmission is probably capable of giving considerable assistance on all kinds of echoes, but can by no means hope to shut out completely the short time echoes referred to in this paper. In fact, this type of scattering by ground reflection may be largely responsible for the widening of the angle of the beam to greater values than would be expected from an observation of its local performance. If ground reflections of this magnitude occur at all they can throw off a great deal of radiation from a beam as well as throw it directly back upon itself.

(6) The hours of the day and the seasons of the year when round-the-world signals can be expected can be predicted for any given pair of stations by reference to the simple principle that round-the-world signal must make most of its night transit in the summer hemisphere and if the great circle between two stations is such that this is not possible, round-the-world echoes will not occur. There is entirely too little information at hand to predict in the matter of the short time echoes. It is urgently requested that observations on a quantitative basis be carried out in various parts of the world on this matter in order to demonstrate whether this is a phenomenon more or less pe-

culiar to this part of the world or is more general in its occurrence. Only by such observations can any decision be made between possible tentative explanations of the phenomena involved.

(7) There appears to be nothing in the nature of echo signals so far observed which is in contradiction with the general theoretical principles already laid down.

THE STATUS OF FREQUENCY STANDARDIZATION*

By

J. H. Dellinger

(Chief of Radio Section, Bureau of Standards, Washington, D. C.)

Summary—*The measurement of frequency, hitherto of laboratory interest only, has become of first-rank importance in reducing radio interference. This has come about through the increasing use of all available radio channels particularly at broadcasting and higher frequencies. While an accuracy of one half per cent was satisfactory five years ago, accuracies a thousand times as good are now sought.*

The piezo oscillator is meeting the needs of this situation in large part. Much effort is being devoted to making the piezo oscillator as constant as possible. Commercially available piezo oscillators, without temperature control, are generally reliable to about 0.03 per cent, just barely enough to meet the Federal Radio Commission's requirement of one-half kilocycle. In order to reach greater accuracy, considerable work is being done on the primary standards of frequency, to insure the highest constancy and accuracy. The Bureau of Standards and other organizations are engaged on a cooperative program to attain an accuracy of 0.001 per cent. Comparisons with other nations show that the national laboratories of the larger countries are already in agreement to about 0.003 per cent.

Temperature controlled piezo oscillators will probably allow the holding of station frequencies so close that several stations can broadcast on the same frequency without heterodyne interference. Use of these or equivalent devices is vital to the maximum utilization of the very high frequencies; the separation of 0.1 per cent between high-frequency stations which is practicable in the immediate future is largely determined by frequency variations, and can be reduced as practice improves.

Introduction

RADIO manufacturers, transmitting stations, and standardization agencies have found it necessary to increase their accuracy of frequency measurement progressively during the past four or five years. This need has come about through the increasing use of the available radio channels and particularly through the development of broadcasting and of high-frequency transmission. While an accuracy of one-half per cent was satisfactory five years ago, it is now necessary to give consideration to accuracies a thousand times as good. It is not merely a question of measurement. Frequencies of transmitting stations

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must be actually held constant with very great accuracy. This is becoming more and more important as the available radio channels become saturated. The maximum number of communications can be packed into the radio spectrum only if each stays within its own channel, as any wandering due to inaccurate frequency adjustment causes interference with the communication on the adjacent channel.

Interference due to frequency variations is perhaps now the main source of interference, as far as technical (apparatus) causes of interference are concerned. Receiving sets are now so well designed that other sources of interference can be tuned out.

Piezo Oscillators

The advent of the piezo oscillator has met the needs of the situation in part. The device is a readily available standard, in terms of which frequencies may be measured to any desired precision (in the sense of fineness of adjustment). The *accuracy* with which frequencies may be thus made available is the subject to which this discussion is mainly devoted. Since piezo oscillators may be used for frequency comparison with practically unlimited precision, the question of their accuracy depends essentially on their constancy, and on the accuracy and constancy of the basic standards in terms of which they are standardized.

When the piezo oscillator was introduced, two or three years ago, it was quite commonly thought that it was an invariable standard of frequency. It has, however, been found that there are a number of factors which introduce variations in the frequency of the electromotive force produced by a vibrating quartz plate, and also in the frequency of response of a piezo resonator. These factors are the temperature, the particular method and details of the mounting of the quartz plate, the circuit constants of the associated circuits, and the methods of coupling between the circuit in which the quartz plate is inserted and the necessary auxiliary and measuring circuits. It is not true that highly accurate comparisons can be made by sending a quartz plate in the mail from one laboratory to another. The effects of the circuit constants may vary the frequency produced by a piezo oscillator several parts in 1000, and therefore a complete piezo oscillator must be sent for accurate comparisons. If it is desired to make measurements to as great an accuracy as one part in 10,000, it is necessary that the voltages used in the piezo oscillator

be carefully measured and that the circuit arrangements and voltages be always the same in use as under the conditions of original calibration. This has been proved by special experimental trials of the effects of different circuit details on the frequency, by theoretical studies, and by the results of actual trial in practice. To carry the accuracy still further, to one part in 100,000, it is necessary not only to control carefully the circuit constants but also to keep the quartz plate at constant temperature.

CONSTANCY AND ACCURACY OF STANDARDS

In order to secure maximum coherence of the results of frequency measurements, and minimum interference between stations adjusted thereby, it is necessary that all measurements be made in terms of the same primary standard, provided that standard has the two requisite properties, accuracy and constancy. Of these two, constancy is by far the more important, since the outstanding need is that the results of frequency measurements agree with one another, regardless of where or when made. If all should be 0.1 per cent different from the actual absolute value there would be no serious practical consequences. but if different laboratories or stations should have standards differing by 0.1 per cent there would be hopeless confusion and interference. This is not to say that absolute accuracy of the primary standard is unimportant; the highest accuracy must be sought, but this is less urgent than the maintenance of a standard of high constancy.

The official primary standard for the United States is that maintained by the Bureau of Standards. During the past four years the Bureau has steadily improved the constancy and accuracy of this standard by continuous research. In order to have a primary standard adequate to the present and the immediate future, the Bureau determined last year to establish and maintain a standard which should be constant to one part in 100,000, or 0.001 per cent. Such accuracy is difficult to attain in almost any line of physical measurement. The work is being carried on with the cooperation of other Government Departments and the various commercial organizations which have had experience in establishing frequency standards or constructing piezo oscillators of great constancy. Four organizations (the General Electric Company, Westinghouse Company, Navy

Department, and Bureau of Standards) have constructed temperature-controlled piezo oscillators which are intercompared and kept under observation at the Bureau. These are also compared against other standards of the Bureau, including quartz plates, tuning forks, and a special frequency meter. Absolute measurements of their frequencies, against time standards, are made by three organizations, the Navy Department, Bell Telephone Laboratories, and the Bureau of Standards.

The absolute measurement of frequency is a most interesting subject, but in this paper I will only be able to mention it briefly. Since frequency is the reciprocal of time, an absolute frequency measurement is a comparison against a standard of time. Ideally, this should be made as directly as possible against the primary time standard, the rotating earth. In practice, measurements are made against a standard clock, permitting frequency measurements in terms of the mean solar second. The measurements are usually made with the aid of an intermediate device (tuning fork, alternator, or oscillator) which produces an audio frequency, this being measured in terms of the clock. The comparison between the audio frequency and the radio frequency of the standard under measurement is made by a step-up process, such as the use of a harmonic amplifier[1] or cathode-ray oscillographs[2].

The measurement may be made even more directly in terms of the time standard by methods which have recently been worked out to eliminate the intermediate audio-frequency device. Such methods have been devised by the National Physical Laboratory in England, the Naval Electrotechnical Institute in Italy, and by the Bell Telephone Laboratories[3] in the U. S.

International Frequency Comparisons

Since both high- and low-frequency signals can be received over large portions of the world, it is of the greatest urgency not

[1] Frequency Measurement in Electrical Communication.—Horton, Ricker, and Marrison. *Trans.* A.I.E.E., **42**, p. 730; 1923.

A Self-Contained Standard Harmonic Wavemeter.—D. W. Dye. *Phil. Trans.*, Roy. Soc. London, **224**, p. 259; 1924.

Establishment of Radio Standards of Frequency by the Use of a Harmonic Amplifier.—Jolliffe and Hazen. Bureau of Standards Scientific Paper No. 530. 1926.

[2] Primary Radio-Frequency Standardization by Use of the Cathode-Ray Oscillograph.—Hazen and Kenyon. Bureau of Standards Scientific Paper No. 489, 1924.

[3] Precision Determination of Frequency. Horton and Marrison. Proc. I.R.E., **16**, p. 137; February, 1928.

merely that all U. S. radio be on a single frequency basis, but that this be true of the whole world. The international aspect of this is being cared for by the national standardizing laboratories of the various nations through intercomparisons of frequency standards. The Bureau of Standards has been particularly active in initiating and directing this work.

International comparisons of any physical quantity can be made by sending the standards of that quantity successively to the laboratories of the different nations. In the frequency comparisons that method has been followed with one exception. About five years ago it appeared that greater accuracy might possibly be obtained by another method, namely, the simultaneous measurement in the various laboratories of the frequencies of waves transmitted from stations of sufficient power to be received simultaneously in many nations. In 1924 such a set of comparisons was organized by the Bureau of Standards and measurements were made in the national laboratories of England, France, Italy, Germany, and the United States. The results of such comparisons extending over a number of weeks were only fairly satisfactory. The accuracy of measurement was not as good as had been hoped. The result, briefly, was to show an agreement between the various countries within about two parts in 1000. The methods used, and independent evidence of various kinds, were such that it was certain that the standards used in each of these countries were more accurate than this. The observed differences were therefore due in large part to the methods of measurement.

From the foregoing experience it was concluded that the more common method (actual transportation of standards from one laboratory to another) used in connection with other physical quantities should be considered anew. Just at that time, about four years ago, enough had been learned of the possibilities of piezo oscillators and piezo resonators to make it appear that they offered particular advantages as portable standards. Measurements upon piezo-electric devices therefore offered a method which might be superior to the scheme of measuring simultaneously the frequencies of transmitted waves. The first attempt of this kind was an informal series of measurements made in several national laboratories by W. G. Cady, Professor at Wesleyan University, Middletown, Connecticut. He took a number of piezo resonators (not oscillators) abroad; the results

TABLE I

Piezo Oscillator B. S. 33465–D with Quartz Plate No. 15

Laboratory	Approx. date	Temperature deg. C.	Frequency	Minus Mean B.S.	Frequency	Minus Mean B.S.	Frequency	Minus Mean B.S.
Bureau of Standards	Dec. 1925	—	75.30	—	106.24	—	455.40	—
National Physical Laboratory, England	Feb. 1926	17.2	75.344	+0.038%	106.273	+0.031%	455.560	+0.016%
Telegraphie Militaire, France	July 1926	(22)	75.33	+0.020%	106.1	−0.132%	455.7	+0.047%
Istituto Elettrotecnico e Radiotelegrafico, Italy	Sept. 1926	26.5	75.347	+0.042%	106.200	−0.038%	455.474	−0.002%
Physikalisch-Technische Reichsanstalt, Germany	Feb. 1927	20.5	75.331	+0.021%	106.248	+0.008%	455.525	+0.009%
Bureau of Standards	July 1927	24.8	75.33	—	106.24	—	455.57	—
Bureau of Standards	Mean B.S.	(24.8)	75.315		106.240		455.485	

TABLE II

Piezo Oscillator B. S. 33465-C with Quartz Plate No. 16

Laboratory	Approx. date	Temperature deg. C.	Frequency	Minus Mean B.S.	Frequency	Minus Mean B.S.	Frequency	Minus Mean B.S.
Bureau of Standards	May 1926	—	75.037	—	105.870	—	455.833	—
Physikalisch-Technische Reichsanstalt, Germany	Dec. 1926	20.5	75.035	−0.009%	105.879	+0.007%	455.940	+0.023%
Istituto Elettrotecnico e Radiotelegrafico, Italy	Mar. 1927	16.5	75.090	+0.064%	105.952	+0.075%	456.152	+0.069%
Telegraphie Militaire, France	June 1927	(22.)	75.004	−0.051%	—	—	455.940	+0.023%
National Physical Laboratory, England	July 1927	19.3	75.054	+0.016%	105.891	+0.018%	455.930	+0.021%
Bureau of Standards	Dec. 1927	22.6	75.048	—	105.875	—	455.840	—
Bureau of Standards	Mean B.S.	(22.6)	75.042		105.872		455.836	

of his measurements are published in his article, "An International Comparison of Radio Wavelength Standards by Means of Piezo-electric Resonators."[4]

In December, 1925, the Bureau of Standards sent a piezo oscillator to Europe, sending it first to the National Physical Laboratory of England, which was to send it in turn to the laboratory of the Telegraphie Militaire in France, to the Naval Electrotechnical Institute in Italy, to the Physikalisch-Technische Reichsanstalt in Germany, and then to be returned to the Bureau. A second piezo oscillator was sent in July, 1926, in the other direction, i.e., beginning with the Physikalisch-Technische Reichsanstalt. It is to be noted that complete piezo oscillators were sent, not merely the quartz plates.

The measurements upon these two piezo oscillators, on account of the necessary delays in getting from one laboratory to another, were only completed in December, 1927. The results are given in Tables I and II.

These results show that piezo oscillators, even without temperature control, provide a far more dependable method of making international comparisons of frequency standards than simultaneous measurements of station frequencies. The indicated differences between the various laboratories have a maximum of eight parts in 10,000; the net result is to show an agreement by this method which may be summarized as a few parts in 10,000. It was uncertain as a result of this work whether the observed differences were due to variations in the piezo oscillators used as the means of comparison or to actual differences in the basic standards of the different countries.

The same method was used, viz., the sending of piezo oscillators, to make comparisons between the frequency standards of the United States, Canada, and Japan. Early in 1927 a piezo oscillator was sent to Canada, was there measured by the Radio Service of the Department of Marine, then returned and measured again by the Bureau of Standards. The results were of the same order as those obtained in the European comparisons, and were considered by the Canadian administration as giving a comparison of standards to as great an accuracy as their measurements permitted. The Bureau of Standards sent the same piezo oscillator in March, 1927, to Japan. The average of the Japanese values differs from the Bureau of Standards

[4] Proc. I.R.E., 12, p. 805, Dec. 1924.

TABLE III

PIEZO OSCILLATOR B. S. 33465-E WITH QUARTZ PLATE NO. 12

Laboratory	Approx. date	Temperature deg. C.	Frequency	Minus B. S.	Frequency	Minus B. S.	Frequency	Minus B. S.
Bureau of Standards	Jan. 1927	—	75.32	—	106.26	—	455.86	—
Radio Service, Canada	Feb. 1927	22	75.27	−0.07%	106.16	−0.09%	455.93	+0.02%
Bureau of Standards	Mar. 1927	23.5	75.32	—	106.26	—	455.86	—
Ministry of Communications, Japan	July 1927	24	75.32	0.00%	106.24	−0.02%	455.85	0.00%

NOTE.—This piezo oscillator has not yet returned from Japan.

measurements by only one part in 10,000. The detailed results are given in Table III.

Piezo oscillators with temperature control began to be constructed in 1927, and a great increase of the possible accuracy of the comparison of standards of different laboratories was thus presented. As I was going to Europe in the summer of 1927 for the Bureau of Standards, an opportunity was afforded to use one of these instruments for comparisons with the standards of the national laboratories of Europe. It was hoped that by taking this improved instrument to the European laboratories a frequency comparison could be obtained that would surpass the previous ones in accuracy. I accordingly had the pleasure of making frequency measurements, all of which were very satisfactory, at the National Physical Laboratory in London, at the laboratory of the Telegraphie Militaire under General Ferrie in Paris, at the Italian Navy Laboratory in Livorno, Italy, and at the Physikalisch-Technische Reichsanstalt in Berlin. I wish to acknowledge the splendid cooperation given me in each of these laboratories by my collaborators, Dr. D. W. Dye at the National Physical Laboratory, Prof. R. Jouaust in Paris, Prof. G. Vallauri in Livorno, and Dr. E. Giebe at the Reichsanstalt. The measurements were made in July and August.

The frequencies of the piezo oscillator were measured at the Bureau of Standards before I left the United States and again after my return. The instrument contained two quartz plates in a thermostatically-controlled heated enclosure, the quartz plates being kept at a temperature of 46 degrees C. Instruments were provided to insure that filament and plate voltages were always the same. The frequencies of both quartz plates were approximately 200 kilocycles.

It was interesting to find that in all the laboratories the same general method is used for frequency measurements of high precision. Typically, there are three generators placed in the laboratory about ten feet apart; one is the piezo oscillator under measurement, one is the auxiliary oscillator or heterodyne, and the third is the standard instrument against which the piezo oscillator is to be measured. There is coupled to each of these a single circuit which contains a receiving set with telephone receivers connected to it. The method of measurement is in all cases some variation of the simple procedure of listening for the frequency difference between the auxiliary generator and the

piezo oscillator, and by one process or another reducing this difference to zero, and then adjusting to equality with the auxiliary generator the standard in terms of which the piezo oscillator is being measured (or else determining the difference between the standard and the auxiliary generator).

TABLE IV

SPECIAL TEMPERATURE-CONTROLLED PIEZO OSCILLATOR WITH QUARTZ PLATES *Y* AND *Z*

Laboratory	Date 1927	*Y* Frequency	*Y* Minus Mean B.S.	*Z* Frequency	*Z* Minus Mean B.S.
Bureau of Standards	June 15	200.122	—	200.142	—
National Physical Laboratory, England.	July 14	200.118	0.000%	200.128	−0.006%
Telegraphie Militaire, France	Aug. 4	200.134	+0.008%	200.149	+0.004%
Istituto Elettrotecnico e Radio-telegrafico, Italy.	Aug. 16	200.119	0.000%	200.137	−0.002%
Physikalisch-Technische Reichsanstalt, Germany.	Aug. 31	200.131	+0.006%	200.152	+0.006%
Bureau of Standards	Nov. 16	200.115	—	200.138	—
Bureau of Standards	Mean B.S.	200.118		200.140	

The results, shown in Table IV, justified every hope. The differences between the different laboratories in these measurements ranged from zero to five parts in 100,000, the average of the departures from the mean being three parts in 100,000. This agreement is indeed as good as the degree of certainty of the national standards. It is concluded that by using a portable temperature-controlled piezo oscillator in which the currents through the circuits are always adjusted to the same value, it was possible to get a 10-fold increase of accuracy over that attained with the simpler type of piezo oscillator used in the previous comparisons. This assurance as to the accuracy of the frequency standards available in the larger countries determined the action of the International Radio Conference in October, 1927, on the subject of frequency measurements, as expressed in Article 3 of the General Regulations of the International Radiotelegraph Convention.

The accuracy attained in these comparisons is not by any means the limit attainable. The instrument used in the comparisons was one upon which complete studies had not been made as to temperature equilibrium, frequency lag, and temperature coefficient of the quartz plate. No doubt as time goes on we can improve upon this. On the whole, however, I feel that it has been demonstrated that the several national laboratories are

measuring frequencies with an accuracy satisfactorily in advance of the immediate requirements of radio practice. The standards of frequency of the larger countries agree sufficiently well to insure against interference provided the transmitting stations are accurately adjusted according to their national standards.

Applications in Broadcasting and High Frequencies

The developments in the accurate measurement of frequencies have their principal applications in those portions of the radio spectrum where the congestion of radio traffic is greatest, viz., broadcasting and high frequencies. The Federal Radio Commission requires every broadcasting station to operate within 0.5 kilocycle of its licensed frequency. At a frequency of 1500 kilocycles this means that the frequency must be maintained within 0.03 per cent. There were no means commercially available by which this could be done, a year ago. At the present time, using as a station frequency standard a piezo oscillator without temperature control but carefully operated, the requirement can just be met. To be certain of satisfactory operation, a piezo oscillator should be investigated over the range of temperatures at which it will operate, as sudden large changes of frequency with temperature occur in some of them.

The advent of temperature-controlled piezo oscillators opens up new possibilities. It is apparent, from the experience I have recounted, that these instruments can be relied upon, under conditions of practical use, to 0.003 per cent or better. At a frequency of 1000 kilocycles, this corresponds to a constancy of 30 cycles or better. In other words, two or more broadcasting stations using these devices could operate on the same frequency and remain synchronized so closely that there would be no audible beat note between their carrier frequencies. This is the only practical possibility at present available for eliminating heterodyne interference. Use of this plan may offer a way out of the hitherto insoluble problem of too many broadcasting stations. It is not an easily operated plan; it requires great care in operation of the piezo oscillators at all stations; but it seems to me at present the most practical of the various theoretically possible plans.

The increase in accuracy of frequency measurements has an application of at least equal importance in the use of very high fequencies. Since 0.03 per cent is about the limit of dependa-

bility of piezo oscillators without temperature control, it is not feasible at the present time to use narrower channels than 0.1 per cent. To illustrate, successive channels can be 10 000, 10 010, 10 020 kilocycles, etc. With a commercially available piezo oscillator used as a station frequency standard, the second of these channels could be reliably held within 10 007 and 10 013 kilocycles. Greater deviation than this would bring the transmission dangerously near the adjacent channels and cause interference. Thus, assuming the use of the best commercially available equipment and great care in the operation of stations, high-frequency channels cannot be spaced closer than 0.1 per cent at the present time.

More stations can be operated on the high-frequency waves almost in proportion to the increase in accuracy of the frequency control of transmitting stations. While it is true that there are other factors influencing the width of channel, e.g., selectivity of receiving sets, nevertheless for CW work it is likely that the channel width could be narrowed to something like 0.01 per cent if first-class temperature-controlled piezo oscillators, or apparatus of equivalent accuracy, were universally used in high-frequency stations. To illustrate again, at 10 000 kilocycles the channels would be 1 kilocycle apart, and the frequency of each would be maintained within 300 cycles. CW receiving sets can readily distinguish traffic with the minimum separation provided under such operation. For the whole frequency spectrum above 2000 kilocycles, this would mean that, in place of some 2000 CW station assignments possible under present conditions, about 20 000 could be accommodated.

Discussions

Henry Shore†: As Dr. Dellinger has pointed out, the second is the logical basis of our frequency measurement and standardization. Since this is so, it seems to me that any frequency standard should be directly and intimately related to the basis. Thus the frequency standard should be controlled by the clock or pendulum. A method of accomplishing this has been suggested by Dr. V. Bush of Massachusetts Institute of Technology.

The method consists of series of multivibrators interlocked and controlled by impulses from the standard time piece, the

† Research Engineer, Radio Corporation of America, New York City. Original Manuscript Received by the Institute, January 24, 1928.

first multivibrator having a fundamental period of 2 cycles per second. The second multivibrator has a period of 40 cycles per second, the third a period of 1600 cycles per second and so on up the frequency spectrum. The clock impulses, furnished say by means of photocell, simply serve to maintain the fundamental frequency constant and do not drive the multivibrator units.

For a more detailed description of the multivibrator the excellent paper of J. K. Clapp, "Universal Frequency Standardization from a Single Frequency Standard," *Journal* Optical Society of America, 15, No. 1, July, 1927, should be consulted.

G. W. Kenrick‡: Dr. Dellinger, in his interesting paper, has directed attention to the importance of the frequency standardization of broadcasting stations and suggested the use of standardized crystals at the stations to insure the constancy of the emitted frequency. I would like to direct attention to another possibility which suggests itself for holding these frequencies at such values as to insure suitable separations.

According to this plan as many stations as necessary would be established as standardization stations to transmit, on a radio channel, a standard frequency preferably equal to the desired frequency separation of the broadcasting stations or some exact sub-multiple thereof (let us say 10 kc. for example). These signals could now be received at the various broadcasting stations; and, after demodulation, passed through sufficient stages of frequency multiplication to bring them into correspondence with the given broadcasting station's assigned frequency. Such frequency multipliers utilizing accentuated harmonics in vacuum-tube circuits and resonance phenomena are, of course, quite well-known to the present stage of the art. Prime harmonics, not readily obtainable by several stages of successive multiplication alone, are readily available by the introduction of sum and difference frequencies produced by the modulation of the output of the higher multiplier stages by that of the lower stages or the fundamental.

It will be noted that this method of standardization has the advantage that the requisite frequency separations are secured independently of the precision of the 10 kc. standard which, however, is a single standard and hence readily producible with a high degree of precision and constancy at some central point,

‡ Instructor, Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia, Penna. * Original Manuscript Received by the Institute, January 30, 1928.

such as the Bureau of Standards. In my opinion this method has a distinct advantage over the use of a large number of independently calibrated crystals maintained at broadcasting stations under conditions necessarily less rigorous than those to which a central standard could be subjected. It will be noted that a minute injury or disturbance in any of these secondary crystal standards will correspondingly produce interference, and not infrequent recalibrations would hence be desirable.

The production of suitable frequency multiplying sets for use in the method here described, particularly in the considerable quantity necessary for the equipment of all broadcasting stations, would seemingly be readily practicable at a per unit cost inconsequential in comparison to the total investment in even a small modern station. It will be particularly noted that, once adjusted to the proper harmonic, the standardization at each station (by dead beat methods) may be obtained without further adjustment of the apparatus and will be essentially independent of local conditions.

J. H. Dellinger: The work mentioned by Mr. Shore is an interesting addition to the work I mentioned on direct comparisons of radio frequencies with a time standard. I judge that the method is the one developed by the National Physical Laboratory. The use of multivibrators is not a good feature, as they are inferior to the harmonic amplifier.

The method Mr. Kenrick mentioned of holding broadcasting station frequencies on the licensed values is theoretically correct. However, outside of the difficulty of distributing the basic frequency to all stations, the most serious objection to the method is the requirement that every broadcasting station use a harmonic amplifier to compare its frequency with the lower frequency standard. This would introduce a complicated apparatus of laboratory type into every station; it is very unlikely that the station personnel would in all cases be competent to secure the desired results.

DETECTION BY GRID RECTIFICTION WITH THE HIGH-VACUUM TRIODE*

By

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Summary—*The theory of detection of small signals by grid rectification with the high-vacuum triode of the 201A type is briefly discussed by means of the mathematical method of Carson. Special attention is given to the grid-leak, grid-condenser arrangement. The relation between the non-linear distortion and the degree of modulation and between the frequency distortion and the grid impedances is discussed. A convenient method is described for experimentally ascertaining the frequency distortion in detection and illustrated by means of the results in a typical case; this distortion is compared with that due to resonance in the r.f. amplifier circuits. A method of securing efficient grid-rectification in the super-heterodyne system is described. The detection coefficients for the 200A alkali-vapor tube are given in an appendix.*

IN an important paper recently published in the PROCEEDINGS (15, 113, 1927) Chaffee and Browning have considered in a very general way the theory of detection of small signals with the three-electrode tube. In the introduction to this paper Professor Chaffee kindly refers to some unpublished manuscript notes which I sent him in 1924 dealing with the special case of grid-circuit detection with the high-vacuum device. In these notes I had made a somewhat different mathematical approach to the subject, which appeared to me to have certain advantages of conciseness and to exhibit the results in a simple and physically fundamental way. I had also restricted myself to the detection by grid rectification because technically it was of most importance. Believing that these old notes might be found to cover some detailed aspects of the grid-detection process not specifically covered in some of the other papers I have ventured in the present paper to reproduce them in abridged form, with the omission of mathematical details.

With grid-circuit rectification the desired result of detection of a modulated carrier-wave is the production at the grid-filament terminals of the tube of a low-frequency voltage faithfully resembling the original modulation. This audio voltage is then amplified by the tube in the usual way. Due to the independence of e_p and i_g the functions of rectification and

* Original Manuscript Received by the Institute, November 11, 1927.

amplification are divorced from each other and the device may be regarded as playing the dual role of an amplifying-rectifier.[1] The theory of the amplification of the replica modulation voltage is well-known and the calculations are carried through by the customary application of the fundamental plate-circuit theorem, taking into account such impedances at the modulation frequency that may be present in the plate circuit.

In the case of plate-circuit rectification the device may also be regarded as an amplifying rectifier, but the amplification now takes place at the carrier frequency and previous to rectification in the plate circuit. The simplicity of this view is somewhat disturbed by the variability of the amplification-factor μ, which should be taken into consideration in calculating the higher order effects.

1. Summary of Assumptions.—The restricted scope of the discussion may be indicated by the following list of assumptions, which are assembled here for convenient reference:

(1) Discussion restricted to high-vacuum triode of the 201A type in the characteristic of which there are no kinks due to excitation of resonance radiation, or to ionization;

(2) The electron grid current is a function only of the grid potential and is not affected by the plate potential;

(3) The intensity of the signals is so small that terms of higher order than the second can be neglected;

(4) Specimen signal is of the type $e_0 = E_0(1-m \sin at) \sin \omega t$.

(5) Plate rectification is negligible compared to that in the grid circuit.

Regarding the accuracy of (1) and (2) see the curves for the 201A tube shown in Fig. 4.

2. Grid-Circuit Rectification. Elements of the Mathematical Problem.—The theoretical elements of the problem are shown in Fig. 1. The impedance denoted by Z_c is commonly called the "grid-impedance" and is in series between the supply network (terminals AB) and the grid circuit of the tube. This is not restricted to the conventional form, i.e., condenser and grid-leak resistor in parallel, but may be of any linear type. The impedance denoted by Z_i (or admittance A_i) represents the input admittance of the tube in so far as it arises from the

[1] This was noticed in 1918 by L. M. Hull, to whom the expressive term *amplifying-rectifier* is due. See "Operation of an Electron Tube as an Amplifying Rectifier," *Phys. Rev.*, 15, 557, 1920.

capacity between grid-filament electrodes, and the feed-back from the plate circuit through the grid-plate capacity. The admittance due to the electron current is in parallel with the input admittance and is non-linear, the functional relationship being denoted by $i_g = f(e_g)$.

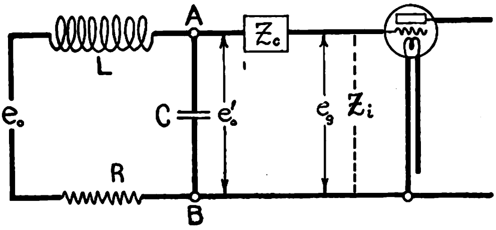

Fig. 1—Elementary Detector and Impact Circuit.

The supply circuit is arbitrarily represented as a simple resonant *LC* circuit, but the restriction is merely one of convenience. As I pointed out in a discussion of a paper before the Institute in 1918,[2] it is necessary in general to take the impedance of this network into consideration. The impressed voltage e_0 may be reduced to an equivalent voltage e_0 across *AB*, and where the impedance Z_c would enter in a discussion which ignored the supply circuit we shall find it necessary to add to Z_c the impedance of the supply circuit looking into it from the terminals *AB*. With these artificialities the problem is reduced to that of a non-linear circuit containing linear and non-linear impedances in series with an impressed voltage. The impressed voltage will be considered to be a modulated carrier of the form:

$$e_0 = E_0(1 - m \sin at)\sin \omega t, \tag{1}$$

where k is the coefficient of modulation. The modulation frequency a will be considered as lying below the carrier frequency ω.

Various mathematical devices are available for the solution of such non-linear problems. In particular a very fundamental

[2] Meeting of November 29, 1918; PROC. I.R.E., 7, 185, 1919. That the impedance of the supply network affects the first-order voltage passed on to the grid circuit of the tube is obvious, and trivial. What I refer to is the effect of this impedance upon the higher-order r.f. voltages, 2ω, 3ω, $\cdots$, $n\omega$ which are generated by the non-linearity of the grid-current characteristic. If the supply network contains impedances which are significant at these frequencies the low-frequency detection terms may be profoundly modified. This may augment or diminish the distortion depending upon the signs and values of the higher-order parameters.

and convenient formulation in an integral equation has been devised by Carson.[8] In the steady state Carson's integral equation solution degenerates into the older power series solution which is somewhat better known and easier to handle. It is especially convenient when small amplitudes are involved and the parameters are such as to insure convergence of the series. Carson has made use of the power series method in discussing the steady state theory of the triode under van der Bijl's assumption of constant μ. I have applied this method to the present problem and the results, so far as the second-order effects are concerned, may be summarized briefly as follows.

3. Modulated Signal; $e_0 = E_0(1 - m \sin at) \sin \omega t$.—It is sufficiently accurate to assume that that e_g, the voltage on the grid of the tube, is a replica of the voltage e_0 of the above form, and that the r.f. drop in Z_c can be taken care of by saying that $e_g = \theta e'_0$, θ being a complex constant. The second order effects, upon which detection is mainly due in the case of small signals, is calculated as follows: It may be shown that the "equivalent second-order" voltage is:

$$-\frac{P_2}{2A}(e_g{}^2) = -\frac{e_g{}^2}{2A}\,\frac{1}{2}\,\frac{d^2 i_g}{de_g{}^2}, \tag{2}$$

where A is the admittance (total) of the GF terminals of the tube for the components of the above voltage. For the modulated signal (1) this voltage is:

$$-\frac{P_2}{A}\vartheta^2 E_0{}^2 [1 - 2m \underset{\substack{\uparrow \\ \text{detection}}}{\sin at} + m^2 \underset{\substack{\uparrow \\ \text{distortion}}}{\sin^2 at}] \sin^2 \omega t. \tag{3}$$

The first term in the brackets represents an inaudible voltage and is of no interest; the second represents the desired voltage corresponding to detection and is proportional to $m \sin at$; the last, corresponding to the square of this quantity, represents amplitude distortion (of the second-order) which is undesirable. We may note that the distortion depends upon the coefficient of modulation, m. The relative distortion might be expressed by:

$$\frac{\text{Distortion voltage}}{\text{Detection voltage}} = \frac{m}{4}. \tag{4}$$

[8] J. R. Carson, *Phys. Rev.*, 17, 116, 1921; or "Electric Circuit Theory and Operational Calculus," Ch. 10, New York, 1926.

In order to keep the distortion down to tolerable values m must be properly regulated. High-quality telephony by this method thus demands a considerable outlay of power in the carrier. I understand that it is standard practice in the better broadcasting stations to limit m to 20–30 per cent, to which corresponds a relative distortion in detection at the receiver of about 5 per cent. The coefficient m is not, however, a very useful conception when the modulation undergoes the wide fluctuation of practice.

Of the equivalent second-order voltage (3) the second term is the useful demodulation effect. It is of main interest, and the other terms need not be further considered. Retaining only low-frequency components, the e_1 component of the voltage across Z_c and Z_{AB} (to which the voltage across the grid is obviously equal), is:

$$e_g = \frac{1}{2}\,\frac{d^2 i_g}{de_g^2}\,\frac{Z_c'}{1+Z_c'A}\,\vartheta^2 m E_0^2 \sin at \tag{5}$$

Here the impedances or admittances are functions of the frequency a.

4. Detection Coefficient for Second-Order Grid Rectification.—The a-frequency voltage acting upon the grid as a result of the rectification is

$$e(\text{audio}) = \frac{1}{2}\,\frac{d^2 i_s}{de_g^2}\,|Z|\,mE_0^2 \sin(at+\phi), \tag{6}$$

$$E(\text{audio}) = D_g m E_0^2; \tag{7}$$

where

$$Z = \frac{Z_c(a)}{1+Z_s(a)A(a)}\,.$$

D_g is the *detection factor for second-order grid-rectification* and when multiplied by the square of the amplitude of carrier voltage and the modulation coefficient gives the amplitude of the audio voltage. If r.m.s. values are employed:

$$E(\text{r.m.s.}) = \sqrt{2}D_g m E_0(\text{r.m.s.})^2 \tag{8}$$

The detection factor consists of two parts: the first, $1/2\; d^2 i_g/de_g^2$, is a tube parameter and equivalent to an expression which I proposed in 1919[4] to represent the detecting action of the tube;

[4] Proc. I.R.E., 7, 139, 1919.

the second, which may be called the *grid-impedance factor*, represents the effect of the grid apparatus (Z_c) cooperating with the tube. The general requirement for efficient detection is that Z_c shall be large for the modulation frequency and small for the carrier frequency. For high-quality telephony the modulation frequency extends over a considerable range, from about 50 to 10,000 cycles per second, and in order to prevent frequency distortion in the process of detection we have the further requirement that Z_c shall remain as nearly constant as possible over this range. As the audio impedance of Z_c increases without limit and the input admittance A_i vanishes the detection approaches its ultimate value:

$$D_g = \frac{1}{2} \frac{d^2 i_g}{de_g^2} \frac{1}{g_g}. \tag{9}$$

This quantity was proposed by L. A. Hazeltine[5] as a detection factor for grid-rectification, expressing the merit of the tube as a detector. While it is appropriate only in somewhat ideal circumstances, as I urged in this 1919 discussion, it is nevertheless of value as indicating the ultimate possibilities of the tube when and if Z_c can be so selected as to take advantage of them. Hazeltine's factor might be termed the *ultimate detection coefficient* or the *optimal detection coefficient.*

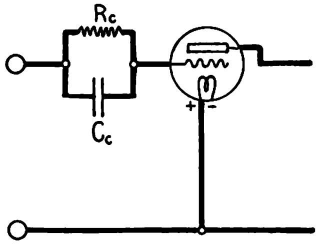

Fig. 2—Grid Leak and Condenser Arrangement for Grid-Circuit Detection.

5. Grid-Condenser and Grid-Leak Arrangement.—With further reference to radio telephony by means of a r.f. carrier it may be of interest to consider briefly the action of the venerable grid-condenser, grid-leak form of Z_c. (Fig. 2) So far as the audio-frequency action is concerned it is a matter of indifference whether R_c is shunted across C_c, as shown in Fig. 2, or connected

[5] L. A. Hazeltine: Proc. I.R.E., 7, 173, 1919.

directly between grid and the positive terminal of the grid-biasing battery E_c, because the audio impedance of the r.f. apparatus LC is negligible. The function of R_c is to provide the high audio impedance necessary for efficient detection, while C_c by-passes the r.f. currents and prevents a significant r.f. drop across R_c. It is interesting to compare this with the older theories in some of the textbooks. In these explanations detection is intimately connected with the charging of C_c, and R_c simply serves in the case of the high-vacuum tube as a leak for this charge, whereas from our present viewpoint R_c is the essential element, and so far as the process of detection is concerned C_c is not only superfluous but harmful as a potential source of frequency distortion.

If ω is large enough the r.f. spectrum of the modulated signal will cover a small range about ω, so that if there be no r.f. impedance in the plate circuit of the detector tube, ϑ is constant. If there is significant carrier-frequency impedance in the plate circuit of the detector, as for example in the case of a resistance-coupled audio amplifier (where the by-pass condenser must be limited in order to preserve reproduction of the high audio frequencies), or in the still more unfavorable case of an R-coupled amplifier used with a superheterodyne receiver with intermediate frequency of the order of 10^5 cycles, the component of input impedance due to feed-back through the grid-plate capacity must be considered and the determination of ϑ can perhaps be made more easily by experiment than by computation.

So far as audio response is concerned the quantity of direct interest is ϑ^2. Fig. 3 illustrates the variation of this quantity in a typical broadcast receiver as the capacity of the grid-condenser C_c is varied. The ordinates are derived from the output audio voltage (with constant modulated r.f. signal in the antenna) which is proportional to ϑ^2. The circuit conditions immediately antecedent to the detector were typical: $L=200$ microhenrys, $C=0.0004$ μfds; 201A tube ($E_b=67$ v., E_c return to positive filament terminal). It will be seen from this curve that the value of C_c could be reduced to about one-half the conventional value (0.00025 μfds.) without much sacrifice of grid voltage.

The grid impedance factor Z in (6) depends upon the grid-electron conductivity (g_g), the admittance of Z_c, and input admittance of the tube (A_i), all at the frequency of modulation. Thus:

$$Z = \frac{1}{g_g + A_i(a) + A_c(a)}. \tag{10}$$

The tube factor g_g may be determined experimentally as a function of the grid-operating voltage. The parameter, dg_g/de_g may then be derived from this data by taking the slope of the conductivity curve for several values of e_g. The admittance of Z_c is easily computed, but the input admittance A_i is somewhat more complicated.

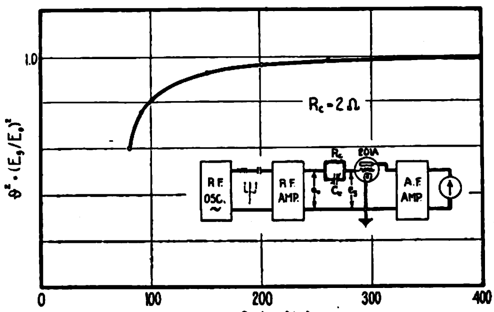

Fig. 3—Variation in $\vartheta^2 = (E_0/Eg)^2$ in a Typical Case as Affected by Grid-Condenser Capacity.

6. Continuation. Special Case; Zero Input Admittance.—To get an idea of the relative effects of these several elements we may temporarily ignore the input admittance, and consider the cooperation of g_g and Z_c. In these circumstances the amplitude of Z is:

$$|Z| = 1/\sqrt{(g_c + g_g)^2 + a^2 c_c^2}. \tag{11}$$

The variation of Z with the modulation frequency a represents a frequency-distortion in the process of detection which is worth noticing in the case of telephone reproduction.

Experimental values of the parameters of a UV-201A type tube which are of interest in the calculation of the detection factor are shown in Fig. 4. This tube was an old one in which the emission had fallen somewhat and is thought to be a better specimen of tubes in use than a new one would have been. The full-line curves represent grid conductance (di_g/de_g) and the dotted curve the grid current i_g for a plate voltage of 45. The actual potential assumed by the grid when using a grid-leak

resistance R_c returned to positive filament ($E_c=5$) are determined graphically as follows:

$$
\begin{aligned}
i_g &= f(e_g) && \text{(a)} \\
i_g &= \frac{E_c - E_g}{R_c} && \text{(b)}
\end{aligned}
\qquad (12)
$$

Curve a is the grid-current, grid-voltage curve of the tube; curve b is a straight line intersecting the E_g axis at E_c and the i_g

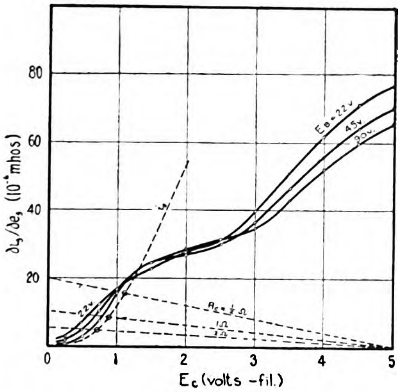

Fig. 4—Experimental Values of Grid Electron Conductance g_g and Grid Current i_g for UV-201A Type Tube Showing Operating Points for Various Values of Grid-Leak Resistance.

axis at E_c/R_c amperes. The intersection of these two curves gives the actual grid potential. The intersections for R_c's of 2, 1, and 1/2 megohms are shown in the figure, amounting to 0.7, 0.9 and 1.1 volts respectively. The values of D_g for various values of R_c as a function of the modulation frequency have been computed from the formula:

$$D_g = \frac{1}{2} \frac{dg_g}{de_g} \frac{1}{\sqrt{(g_g+g_c)^2 + a^2 c_c^2}} \qquad (13)$$

These are shown in Fig. 5.

For low modulation frequencies the detection factor increases with the grid-leak resistance and is approximately equal to unity for 2 megohms. The frequency distortion due to the shunting effect of the grid condenser ($C_c = 0.00025$ μfd.) at the higher frequencies also increases with R_c; for $R_c = 2$ megohms the ampli-

tude at 5000 cycles has fallen to 62 per cent of its normal value. If the higher frequencies are to be preserved it appears desirable to avoid this distortion by employing a low R_c. This involves a double sacrifice in amplitude, double because not only the detection factor is reduced, but the losses introduced into the r.f. LC circuit by the increased electron conductivity further reduce the r.f. amplitudes. From the latter viewpoint it is somewhat preferable to keep the grid-electron conductance low and to obtain the desired high value of $g_g = g_c$ by making g_c high. The

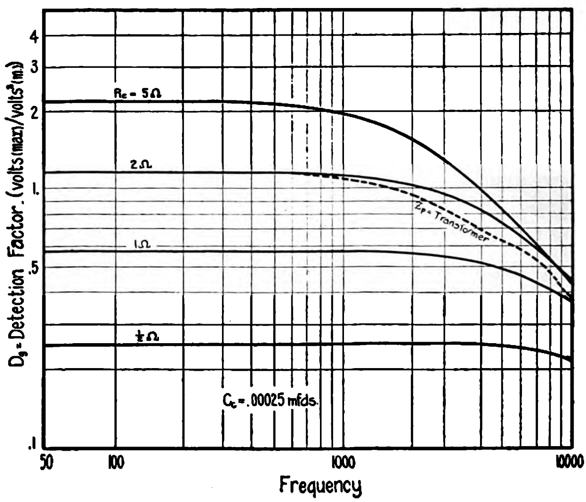

Fig. 5—Detection Coefficient of Second-Order Grid Rectification for UV-201A Tube for Various Grid Leaks as Affected by Audio-Modulation Frequency. Dotted Curve Showing Effect of Audio Input Impedance (trans. plate load).

proper bias for this may be obtained by disconnecting the return from the positive filament and using a battery E_c of less than 5 volts. This will minimize the r.f. losses due to this adjustment, but there still remains the decreased detection factor. The makers of the UV-201A tube recommend a grid-bias of from 5 to 8 megohms. Under these conditions the sensitivity will be good but the frequency distortion will be rather severe.

The above recommendation has been defended by W. B. Roberts of the Radio Corporation Test Department,[6] who calculates that with a 201A tube the use of a 5-megohm leak will

[6] Proc. I.R.E., 15, 793, 1927.

result in a decrease of but 10 per cent at 4500 cycles. The grid resistance at the resulting operating point is stated to be 65,000 ohms. This conclusion is in startling disagreement with the conclusions of this paper (as well as the data of Chaffee and Browning) which is to be attributed to Mr. Roberts' error in choosing the grid resistance. This is actually more nearly 250,000 ohms, and 65,000 ohms corresponds to a grid-leak of 0.5 megohms, not 5 megohms. In these circumstances the frequency distortion would be serious, as Fig. 5 shows. This is confirmed experimentally when the plate rectification is relatively small (as is usually the case), and the impressed voltages lie within the bounds contemplated by the "small signal theory" (i.e. below about 0.3 volt r.m.s.). Both plate rectification and higher impressed voltages tend to offset the frequency distortion (*vide* Sec. 8) and the experimental results under these conditions may be deceptive.

7. Continuation. Effect of the Input Admittance, A_i.—In the preceding section the input admittance was assumed to be zero in order to exhibit the action of the grid-condenser and grid-leak in the simplest possible way. We may now take this into consideration and briefly indicate its effect.

The input admittance in question is that presented at the modulation frequency a. As occasioned by the grid-plate capacity of the detector tube it will depend among other things upon the audio impedance in the plate circuit. Two types of plate load are of practical interest: (1) an audio interstage coupling transformer, and (2) a coupling resistor; both will usually be shunted by a small by-pass condenser to reduce the r.f. input admittance. If the precise nature of these impedances is known over the audio range the input admittance may be computed from formulas previously published;[6] but this is likely to prove a vexatious process and the required values are perhaps more easily arrived at by direct measurements. The results of such measurement of the components g_i (input conductance) and C_i (input capacity) for the UV-201A tube are reproduced in Fig. 6a for the two types of plate load.

In the first case, the transformer (R.F.L. Model 3A) was of average characteristics as follows: $L_1=40$ h., $L_2=430$ h., $\alpha L_1=0.4$ h., $C_2=0.00016\mu$fd., ratio 3.3/1. The tube constants were as follows: $E_b=45$ v., $E_c=1$ v., $\mu=8$, C_m (grid-plate) $=8.5$

[6] J. M. Miller, "Scientific Papers of the Bureau of Standards," No 351, 1919. Stuart Ballantine: *Phys. Rev.*, 15, 409, 1920.

$\mu\mu$fds., C_1 (grid-filament) $= 4$ $\mu\mu$fds. The referred secondary capacity antiresonates with L_1 at about 700 cycles, marked n_0 on the curve. Below this the transformer reactance is inductive and g_i is negative. Above n_0 and up to frequency n_1, near which the referred secondary capacity resonates with the leakage reactance, the reactance is negative and g_i is positive. Above n_i the leakage reactance prevails for an interval in which g_i again becomes negative. The input capacity, C_i, varies from about 112 $\mu\mu$fds. to a minimum of 33 $\mu\mu$fds. near n_1.

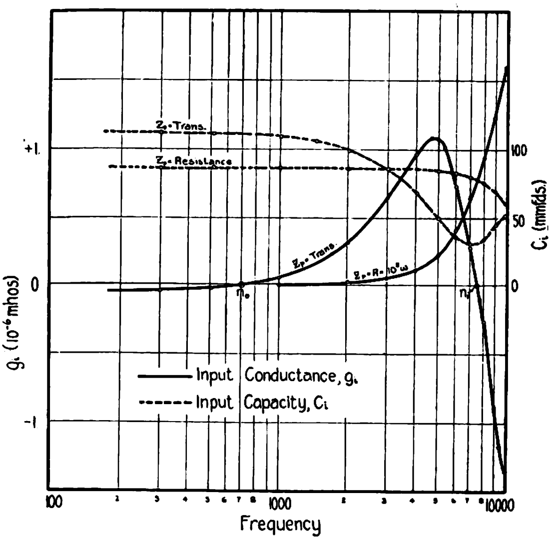

Fig. 6a—Input Conductance, g_i, and Input Capacity, C_i, of UV-201A Tube as a Function of Audio-Frequency for Two Types of Plate Load, (1) Transformer and (2) 100,000-ohm Coupling Resistor.

In the case of the R-coupling arrangement, $E_b = 90$ v., and the coupling resistor was 100,000 ohms with a by-pass condenser of 0.0005 μfd. Here g_i rises and C_i falls continuously with increasing frequency.

The effect of these input admittance variations upon the "impedance factor" Z and the detection factor D_g may be computed from (10). The results are represented by the dotted curve in Fig. 5 for the case of a 2-megohm grid-leak resistor and an audio-transformer plate load. The effect of A_i is not particularly striking. The chief effect is that due to the increased total capacity, which has been increased by C_i about 50 per cent.

The curves for the transformer and resistor will therefore be very much alike. In the absence of A_i the amplitude had decreased to about 62 per cent at 5000 cycles; with A_i it is decreased to about 50 per cent of its low-frequency value.

By arranging a proper audio-frequency feed-back I have found it possible to compensate the falling off of the detection factor at high audio frequencies, and even to give it the rising characteristic which is useful in compensating the frequency distortion due to the selectivity of tuned circuits, audio-amplifier and loudspeaker characteristics, etc.

8. Effect of Plate Rectification upon Frequency Distortion.—For small signals the rectification in the plate circuit may be represented by an equivalent voltage, E, acting as usual through the plate resistance and external impedance, where:

$$E = D_p E_0^2 m \vartheta_p^2 \ ; \ D_p = -\frac{1}{2} \frac{\delta^2 i_p}{\delta e_g^2} \frac{1}{g_p}$$

and $\vartheta_p = Z_p(\omega)/Z_p(\omega) + R_p$. The quantity ϑ_p depends upon the degree of by-passing of the external impedance in the plate cir-

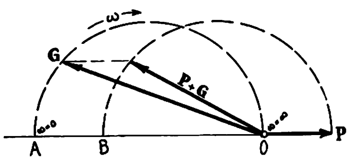

Fig. 6b—Vector Addition of Grid and Plate Rectification; Grid-Leak and Grid-Condenser Arrangement.

cuit for the r.f. components and is analogous to the corresponding quantity for the grid circuit. Considering the grid-condenser, grid-leak arrangement and adding the effects of plate rectification to those due to the grid, we get for the total voltage effective in the plate circuit:

$$E = E_0^2 m \left[D_p \vartheta_p^2 - \frac{1}{2} \frac{\delta^2 i_g}{\delta e_g^2} \frac{1}{g_c + g_g + A(a) + A_c(a)} \mu \right]. \tag{14}$$

$\uparrow$ plate rectification $\qquad$ $\uparrow$ grid rectification

The combination of these two rectification effects is illustrated vectorially in Fig. 6b. The vector G represents the grid detection; as the frequency varies from 0 to infinity its endpoint describes the semicircle AO. The plate detection is represented by the vector P which is constant with respect to frequency. The vector sum of these revolves about the point O and its end point describes the semicircle BP. It is immediately clear that the effect of plate detection is to reduce both the total detection and its variation with frequency. In particular when $P=1/2\ G$ (zero frequency), or:

$$D_p\delta_p{}^2=\frac{\mu}{4}\ \frac{\delta^2 i_g}{\delta e_g{}^2}\ \frac{1}{g_c+r_g},$$

the frequency distortion vanishes and the detection is halved. When $P>G(O)$ the detection increases with the frequency instead of diminishing. The experimental curves shown in Fig. 6c

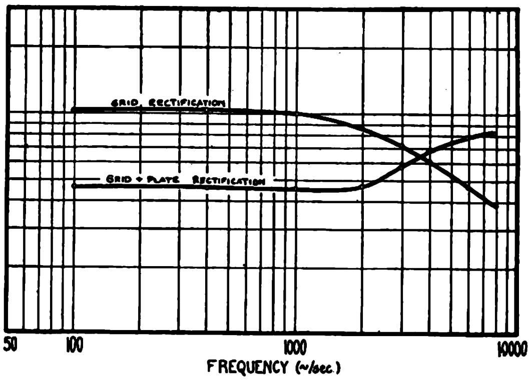

Fig. 6c—Effect of Plate Rectification on Frequency Distortion; Nominal Grid Detection with Grid-Leak and Grid-Condenser; 201A Tube, 5,000 meters.

illustrate the two types of variation. For conventional values of the constants grid rectification is so much more efficient than plate rectification that the latter affects the distortion to a negligible extent.

This furnishes another method of adjusting the distortion; like that suggested in Section 6 (i.e., reducing $C_c\ /g_c+g_g$) it involves a sacrifice in efficiency.

9. Experimental Method for Determining Frequency Distortion.—The frequency distortion in a grid-circuit detector caused by the variation of the grid impedance factor may also be investigated directly by experimental methods. Fig. 7 illustrates a method[7] which I have used successfully for several years and as applied to the usual grid-leak, grid-condenser arrangement have found simple and convenient when a calibrated modulated r.f. carrier is not at hand. The constant voltage furnished by an adjustable audio oscillator is introduced in series with the grid-leak R_c as shown. If not normally so connected the grid-leak is connected between grid and filament during these measurements to permit grounding the a.f. source. The audio voltage is kept small enough so that first-order effects alone are concerned; it is then not necessary to filter the output before leading it to

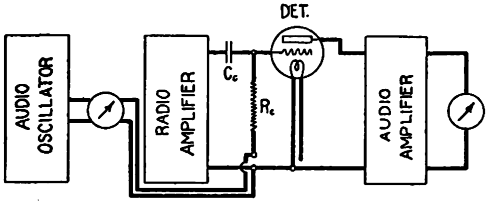

Fig. 7—Experimental Method of Studying Frequency Distortion in Grid Detection (Grid Leak and Grid Condenser) using Audio-Frequency Oscillator.

the output meter. The "amplifier" should preferably have a gain at least equal to that of the audio amplifier of the average radio broadcast receiver. If the frequency-response curve for the amplifier (including the amplifying performance of the detector tube) is not known it may be found at the same time by making $R_c=0$ and plotting output against frequency. Absolute values are of no interest. The ratio of the two curves found with $R_c=0$ and $R_c=R_c$, corrected so that it approaches unity at low frequencies, varies with frequency in precisely the same way that the grid impedance factor varies with frequency. For if e represents the impressed audio voltage, then the voltage e_g across the grid-filament terminals will be

$$\frac{e_g}{e}=\frac{g_c}{g_c+A+i\omega c_c}=g_c\times\text{grid } z \text{ factor.} \tag{15}$$

[7] This scheme was devised independently by H. A. Wheeler; (Discussion at I.R.E. Convention, January, 1927.)

The convenience of the method depends upon the grid-leak impedance being a pure resistance; in the general case other methods must be resorted to.

It should be pointed out that this method measures only the variation of grid impedance and will not give a true picture of the frequency distortion when there is appreciable plate rectification. If plate rectification is suspected it is preferable to use for the measurements a calibrated r.f. carrier, modulated at various audio frequencies.

10. Relation of Detector Distortion to other Sources of Frequency Distortion in a Typical Receiver.—A question of practical interest is the relation of the frequency distortion in the detector to other sources of frequency distortion in the receiver.

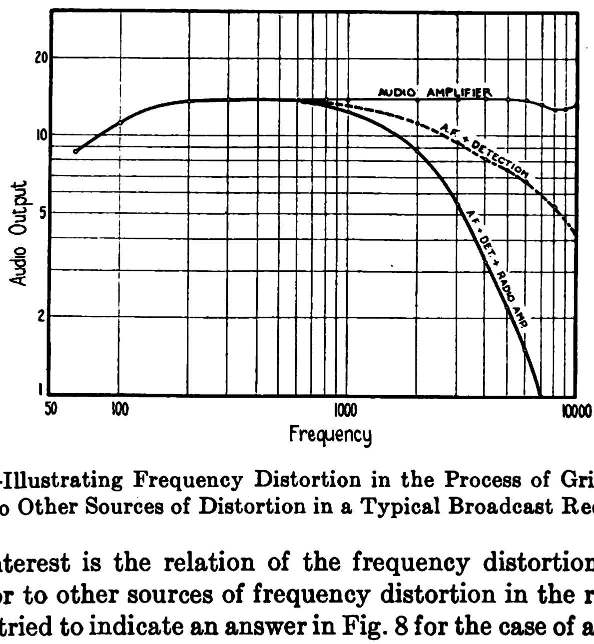

Fig. 8—Illustrating Frequency Distortion in the Process of Grid Detection to Other Sources of Distortion in a Typical Broadcast Receiver.

I have tried to indicate an answer in Fig. 8 for the case of a typical five-tube broadcast receiver. This receiver employed two stages of r.f. amplification and three tuned circuits, the circuits being arranged in the form of a Wheatstone a-c. bridge to render each stage monodic so that true cascade selectivity would be obtained. The top curve, labeled *Audio Amplifier*, represents the distortion in the two-stage audio amplifier, the amplification being measured as the ratio of voltage at the grid of the power tube to that at the grid of the detector tube (as described in the last paragraph, $R_c = 0$). The power tube contained its normal electrophone

load but so filtered as to prevent coupling through the internal impedance of the common *B* battery. The transformers (R.F.L., Model 5A) were two which I had designed for another purpose and happened to be at hand. In addition some artificial feed-back was introduced between the stages in order to iron out residual irregularities in the response curve. The resulting response curve is fairly uniform and suitable for this comparison.

The effect of the detector distortion is added to the audio amplifier in the dotted curve, which was taken by the method of the last section. Conditions were typical: 201A tube, $R_c = 2$ megohms, $C_c = 0.00025$ μfds.

The final heavy curve represents the distortion in the entire receiver and was obtained by introducing into the antenna a constant carrier modulated at various audio frequencies by a constant amount (20 per cent). The frequency of the carrier was 750 kc. (400 m.). The distortion due to the selectivity of the r.f. amplifier is rather pronounced at this frequency; at 1300 kc. (230 m.) it is scarcely noticeable up to 5000 cycles. The distortion in the process of detection thus appears to be important in comparison with that due to selectivity in the r.f. amplifier even at a frequency where this latter is most prominent. At 1300 kc. the detector distortion is considerably more important, and tells the whole story if the audio system is good.

11. Application to the Super-Heterodyne.—The superheterodyne utilizes two detectors: the first converts two frequencies of the order of 10^6 into a difference frequency of the order of 10^5; the second converts this difference frequency into audio frequencies ranging from 0 to 10^4. Thus the frequencies in each detector differ by one order of magnitude. In the ordinary receiver they differ by at least two orders of magnitude (carrier 10^6, audio response 10^4). It has already been shown that in the ordinary receiver the use of a grid-condenser, grid-leak form of Z_c is attended with difficulties; in the superheterodyne these difficulties are considerably magnified. In the second detector it is necessary to filter the plate circuit carefully in order to keep ϑ high, particularly when a resistance-coupled audio amplifier is employed. The problem becomes really acute, however, in the first (frequency-changing) detector.

A theory of detection which considers only the second-order effects will not, of course, apply to the process in the first detector of the superheterodyne. If this were the case the signal amplitude

would rise continuously with the heterodyne voltage. Actually, as Armstrong showed experimentally, it reaches a maximum and an "optimum heterodyne" exists. Appleton and Taylor have given a rough mathematical explanation of this in terms of the higher order parameters of the tube, of which the fourth seems to be of most importance. However, according to the series analysis of this problem given in Sec. 2, each higher order effect is derived from that of next lower order, and it may be inferred from the second-order terms that in the usual arrangement of the first detector in the superheterodyne the detection takes place by plate-circuit rectification and not by grid-circuit rectification as the frequent connection of a grid-leak and

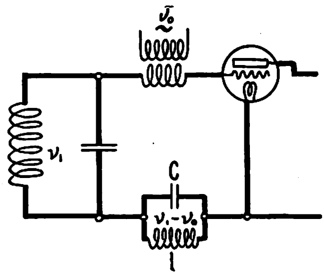

Fig. 9—Arrangement for Increasing the Efficiency of Grid-Rectification in the First Superheterodyne Detector. Anti-resonant Circuit *LC* is Tuned to Difference (intermediate) Frequency.

condenser in the grid circuit by many persons would seem to suggest. Consider the efficiency of the grid-condenser, grid-leak arrangement in this case.

For $C_c = 0.00025\,\mu$fd. and $R_c = 2$ megohms the detection factor at a frequency of 10^4 according to Fig. 4 has fallen to 40 per cent of its d-c. value; at an "intermediate frequency" of 10^5 it will have fallen to 4 per cent of its d-c. value on account of the low impedance of C_c for this frequency. In these circumstances the contribution of the grid-circuit rectification is very doubtful. In the plate circuit, on the other hand, the coupling transformer or filter has a high impedance at the intermediate frequency and efficient detection probably takes place here.

In order to see if the efficiency of detection in the grid-circuit under conditions of optimum heterodyne might not be better than that of the plate circuit under similar conditions I investigated, several years ago, the scheme shown in Fig. 9. Here the

circuit LC, anti-resonant at the intermediate frequency, is placed in series with the grid and replaces the grid-condenser and leak. The constants can be so chosen that the impedance is low at the carrier frequency and maximum at the intermediate frequency. The efficiency of detection in the grid-circuit is thus increased from 4 to about 50 per cent. In order to avoid complications due to feedback from the plate into LC through the grid-plate capacity I used a neutralizing arrangement. Among other things a very noticeable increase in selectivity in the i.f. tuning is to be

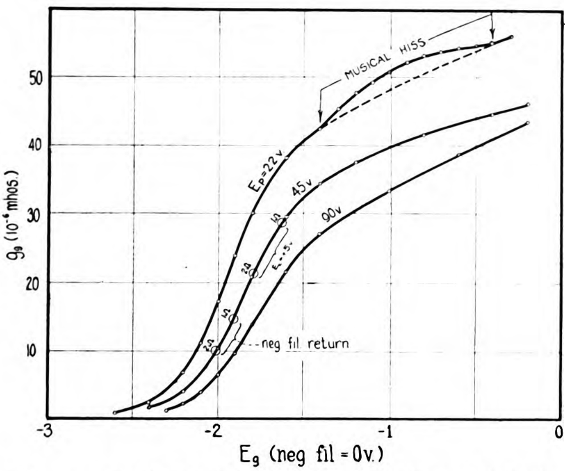

Fig. 10—Electron Grid Conductance of 200A Type Detector Tube for Various Plate Voltages. Small Circles Indicate Operating Points with 1 and 2 Megohm Grid Leaks. Returned to Positive or Negative Filament Terminals.

observed immediately with this arrangement. The same thing can be done in an ordinary detector where sharp audio discrimination is desired, as in the case of c.w. telegraphic work to eliminate noise. In this case LC are tuned to perhaps 1000 cycles. The question as to the relative merits of grid and plate-rectification for frequency changing in the heterodyne is not relevant to the present paper and will be considered in detail in a later publication. Fig. 8 is included merely to indicate how grid rectification may be obtained, if desired.

Addendum

Note on the Detection Characteristics of the 200A Type Tube.—Since this paper was written the 200A type detector

tube has been placed upon the market. This tube appears to contain alkali vapor, derived from a capsule contained in a small hemispherical boss welded to the plate. So far as I know, very little technical information has been published relating to the theory of operation of this alkali vapor device or to its detection characteristic. It may therefore be of some interest to include curves of grid conductance for a specimen tube selected at random from stock, and a calculation of the detection factor for small signals.

From the viewpoint of detection the main thing of interest is the variation of grid conductance with the grid potential. This is shown in Fig. 10. The grid is of fine mesh and the amplification factor is about 20; this shifts the region of operation to negative values of E_g. The correct negative operating potential may be obtained with the usual values of grid-leak by changing the grid return from the positive to the negative terminal of the filament. The actual operating points for grid-leaks of 1 and 2 megohms for both positive and negative returns are marked on the curve for $E_p=45$ v. It is noticeable that in this tube the plate has more influence upon the grid conductance than was the case with the 201A type tube, hence our separation of the amplifying and detecting functions is not strictly correct. The operating point giving the greatest range of linearity between g_g and E_g is found at $E_g=-1.85$ v., which may be approximated by the use of a 2-megohm leak with positive filament return or a 1-megohm leak with negative filament return. The range of linearity is about 0.1 volt, which in view of the high amplification factor will be adequate to take care of the amplitude of the carrier voltage.

The detection factors for low frequencies at various operating points are given in the following table:

TABLE I

LOW FREQUENCY DETECTION-FACTOR (GRID-RECTIFICATION) FOR 200A TUBE ($E_p=45$ v.) FOR VARIOUS GRID-BIASES

R_c	E_c	E_{g0}	g_g	D_g
1	5 v.	1.62 v.	2.85×10^{6}	0.567
2	5	1.8	2.1	1.4
1	0 v.	1.91	1.5	1.97
2	0	2.0	1.0	2.06

A comparison with the factors for the 201A tube (Fig. 5) will indicate that there is not much difference between D_g for the

two types for the same grid conductance (same r.f. loss); the difference, however, favors the new tube. The increase in signal strength which is observed when this tube is substituted for the 201A is therefore probably to be accounted for by its increased efficiency as an amplifier which results from the much higher amplification factor. The plate resistance at the operating point ($E_g = -1.85$) is about 35,000 ohms. In view of this an audio coupling-transformer designed to operate with the output resistance of the 201A tube will not be satisfactory with the new tube, necessitating redesign with probably a lower turns-ratio. When this is attended to and the same frequency-response characteristic is attained it may be found that the over-all performance has been reduced to about that of the 201A tube. For resistance-coupling the tube may have some advantage, provided the plate circuit is carefully filtered to keep down r.f. input impedance.

It is a pleasure to acknowledge the experimental assistance rendered by Mr. Raymond Asserson.

List of Symbols

$A_g = g_g = 1/R_g$ = Admittance of grid-filament terminals of tube due to electron current.
A_i = Input admittance exclusive of electron current.
$A = A_i + A_g$ = Total admittance of GF terminals of tube.
A_c = Admittance of apparatus in series with grid circuit of tube.
g_i = Conductance component of input admittance.
C_i = Capacity component of input admittance.
C_c = Capacity of grid-condenser.
$g_c = 1/R_c$ = Admittance of grid-leak.
i_g = Electron grid current.
E_0 = Amplitude of carrier voltage.
e_0 = Modulated signal $= E_0(1 - kF(t)) \sin \omega t$.
e = Audio voltage across GF due to detection.
E = Amplitude of above.
m = Coefficient of modulation.
$P_1, P_2, \cdots, P_n$. Differential grid-current parameters of tube.
D_g = Detection factor for grid-circuit rectification.
Z = Grid-impedance factor.
e_0' = Equivalent voltage across AB (Fig. 1) due to e_0.

RECENT DEVELOPMENTS IN LOW POWER AND BROADCASTING TRANSMITTERS

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Summary—Various types of radio transmitting equipments are described, ranging in output from 200 to 2000 watts. The application of master-oscillator power-amplifier circuits for low- and medium-frequency transmitters is explained, and the uses of quartz crystal control for high-frequency and broadcast transmitters are described. A brief explanation is also given of the equi-signal system of radio beacon transmission, which promises to become an important aid in the navigation of aircraft.

THE purpose of this paper is to describe the more important commercial developments which have taken place in low-power transmitting equipment during the past two or three years. The term "low power" as used in this paper refers to equipment with an output of less than 5 kw. This is a more or less natural division of radio transmitting apparatus since outputs of less than 5 kw. are generally obtained with air-cooled vacuum tubes using motor generator sets as a source of power supply, while transmitters having outputs of more than 5 kw. utilize water-cooled tubes with high voltage rectifiers as the usual source of plate supply.

A definition of the method of rating vacuum-tube equipment is essential in comparing various classes of apparatus or in discussing overall efficiency, arrangement of tubes, and general performance. The transmitters covered in this paper have their ratings based on the power delivered to the antenna, exclusive of all losses in loading inductors or antenna series capacitors included within the transmitter proper or mounted externally. This method of rating is of considerable importance in determining the design of a transmitter since unfavorable antenna conditions and other factors often result in high losses in the antenna loading system which are not included as part of the output from the transmitter.

Vacuum-tube transmitter development during the ten years from 1917 to 1927 might be divided roughly into three periods. From 1917 to about 1920 we may say that the fundamental

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oscillator circuits were fairly well-determined and limited military and commercial applications of tube transmitters were made. The problem of designing reliable apparatus required considerable attention during this period, while larger size vacuum tubes were developed to meet the demand for higher outputs.

In the period from 1921 to 1924, we have seen a considerable revision in the type of circuits used in the various transmitters.

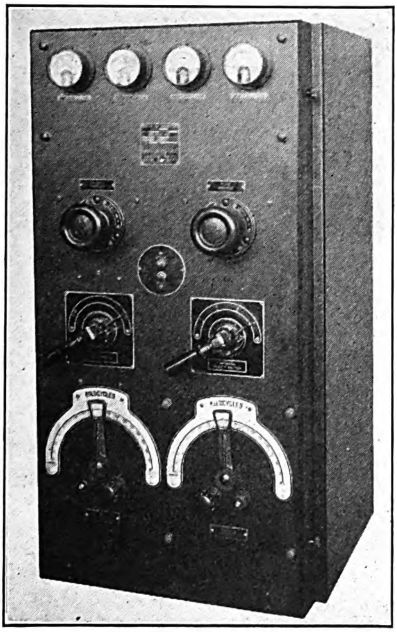

Fig. 1—Front View of Transmitter, Model T-4.

Master-oscillator circuits, providing continuously variable frequency control with a high degree of stability, supplanted the former antenna oscillator circuits. Much progress was also made in securing simplified controls. In the older transmitters, it was necessary to make critical adjustment of tapped coils or coupling devices and transmission was generally limited to a definite number of frequencies. The newer designs permit all adjustments to be made from the front of the panel and the loading on the vacuum tubes is not dependent upon the skill of the operator.

Work done during the past three years or so has consisted largely of commerical development of transmitters suitable for

high-frequency operation, together with refinement in circuit and mechanical design of low- and medium-frequency transmitters. The exacting requirements for constant frequency in the case of the very high frequency equipment have been met through the application of quartz crystal control. The use of crystal control brought about the development of cascade radio-

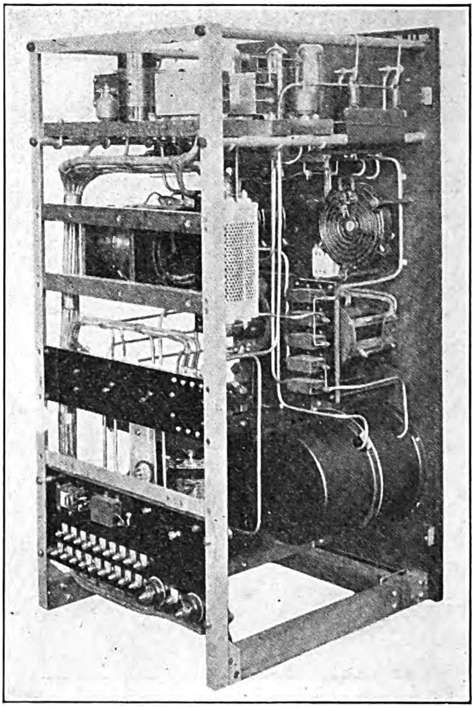

Fig. 2—Rear View of Transmitter, Model T-4.

frequency amplifier circuits so that the comparatively small output obtained from the crystal-controlled tube could be amplified to the desired level. In order to secure high frequencies from comparatively thick crystals, frequency multipliers were also utilized by taking advantage of the harmonics produced in the plate circuit of a tube when the grid is properly biased and supplied with a high exciting voltage. Crystal control has also been applied in the broadcast field so that a very high degree of frequency constancy may be attained by a station. Equipment of this type is described in detail further in this paper.

200-Watt Transmitters

Transmitting equipment, having an output of 200 watts, was developed for the U. S. Coast Guard for service on many of their smaller vessels. A front view of the transmitter, which is known as model *T-4*, is shown in Fig. 1 and a rear view in Fig. 2. Particular attention was directed in the design of this transmitter to reduce the number of controls to a minimum and at

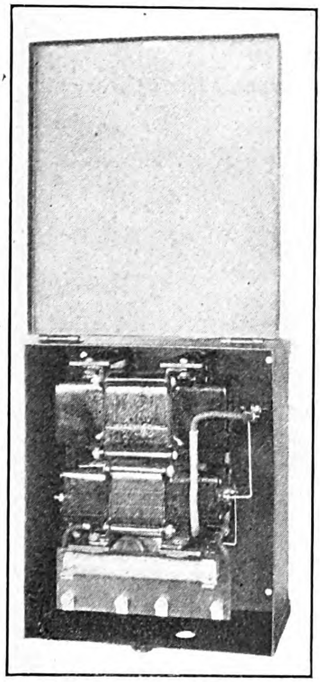

Fig. 3—ICW Telegraph Attachment for Model T-4 Transmitter.

the same time provide for setting the frequency and maintaining it to a high degree of accuracy.

The transmitter utilizes four CG-1984 (UV-211) vacuum tubes, one functioning as the master oscillator and the other three as radio power amplifiers. When used on the average small ship antenna, whose capacitance generally lies within the values of 0.0006 to 0.001 μfd., the transmitter covers a continuous frequency range from 250 to 500 kilocycles. CW and ICW telegraphy are obtained by placing the signal switch on the panel in the proper position. The two large pointers near the bottom

of the panel control the master oscillator variometer and the antenna variometer. The desired transmitting frequency is set by means of the master oscillator variometer and the antenna circuit is then resonated for maximum current by means of the antenna variometer. In addition, the antenna inductance switch is provided in order to select any one of four taps which are provided on the antenna loading inductor.

The four vacuum tubes are mounted in a cushioned cradle near the top of the transmitter in order to permit adequate ventilation and provide for easy replacement. Four instruments are also mounted near the top of the panel and indicate antenna

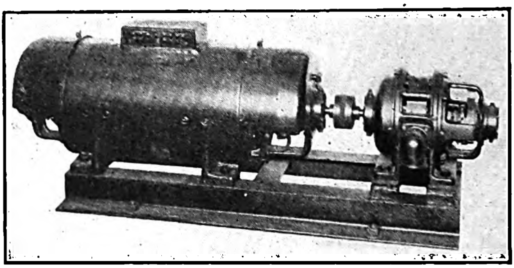

Fig. 4—Three-Unit, Four-Bearing Motor-Generator Set.

current, filament voltage, total plate current, and plate voltage. Rheostats are supplied to permit convenient control of the filament and plate voltages. A momentary contact start-stop push button is mounted in the center of the panel for starting and stopping the motor generator set. A similar switch is also supplied for external mounting on the operator's table so that the equipment may be controlled from either location.

The rear view shown in Fig. 2 illustrates the arrangement of the various component units in the transmitter. The metal container in the lower right section shields the master-oscillator variometer. The capacitors in the master-oscillator circuit are mounted directly above the variometer. All high-frequency wiring in the transmitter is made with copper tubing while low-frequency and control circuits are run in lead-covered wire.

In order to provide ICW telegraphy with this equipment, a separate attachment is supplied as shown in Fig. 3. This unit, whose operation will be described more in detail later, is designed for installation in any convenient place in the radio room and contains no moving parts and requires no adjustment.

A 3-unit, 4-bearing motor-generator set is supplied to furnish filament and plate power and power for the ICW attachment. This machine is shown in Fig. 4, and the automatic starter for controlling it, in Fig. 5. The motor operates from a line supply of 110 to 125 volts direct current and is also built for operation from a 32-volt d-c. system. Slip rings are provided on the motor

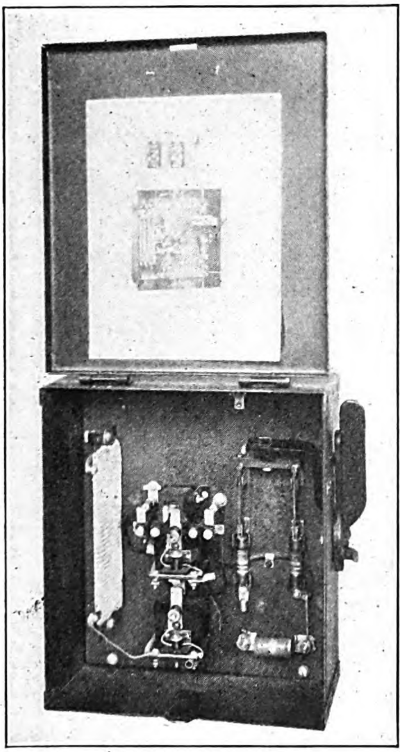

Fig. 5—Automatic Starter.

winding in order to furnish alternating current for heating the filaments of the vacuum tubes. The plate generator is rated 0.8 kw. at 1000 volts direct current. The third unit of the M.G. set is known as the tone alternator and delivers 0.25 KVA at 110 volts and 500 cycles. Ball bearings are provided in this motor generator set.

The theory of operation of this 200-watt transmitter may be understood by referring to the schematic circuit diagram in Fig. 6. The master-oscillator tube, consisting of one CG-1984 tube, has its frequency controlled by the variometer *L*-4 which

operates in a capacity-coupled circuit. This variometer is shunted by three capacitors connected in series marked C-1, C-2 and C-4. Capacitor C-3 is the usual plate-blocking capacitor and serves to keep the d-c. plate voltage off the oscillating circuit and at the same time provides a low reactance path for the radio-frequency energy. The d-c. plate supply for the master-oscillator tube is obtained from the 1000-volt generator through the radio

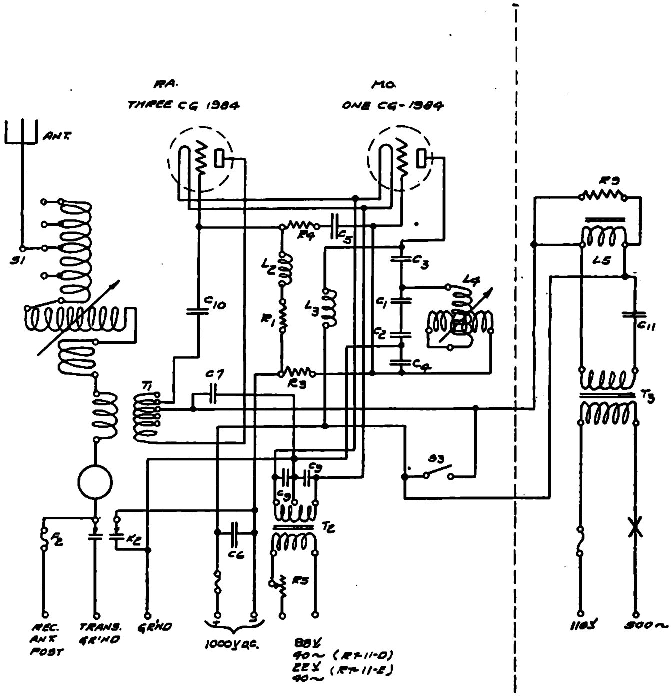

Fig. 6—Schematic Circuit Diagram of Model T-4 Transmitter.

choke L-3. Grid excitation to the three CG-1984 power amplifier tubes is obtained through capacitor C-5 and resistor R-4. The grid-leak circuit of the radio-amplifier tubes is made up of inductor L-2 which acts as a radio-frequency choke, and resistor R-1. The grid leak on the master-oscillator tube is resistor R-3.

Capacitor C-10 is used to stabilize the radio amplifiers, while capacitor C-7 is the plate by-pass unit and provides a low reactance path for the high-frequency energy in the power-amplifier plate circuit.

The power output from the radio-amplifier tubes is delivered to the antenna through the antenna transformer T-1. Flexible

leads are provided to taps on the primary of this transformer to make adjustments for unusual antenna conditions. In the ordinary case, these adjustments are made at the factory and it is not necessary to readjust them in the actual installation.

The antenna circuit is adjusted for resonance to the frequency of the master oscillator by means of the antenna loading inductor

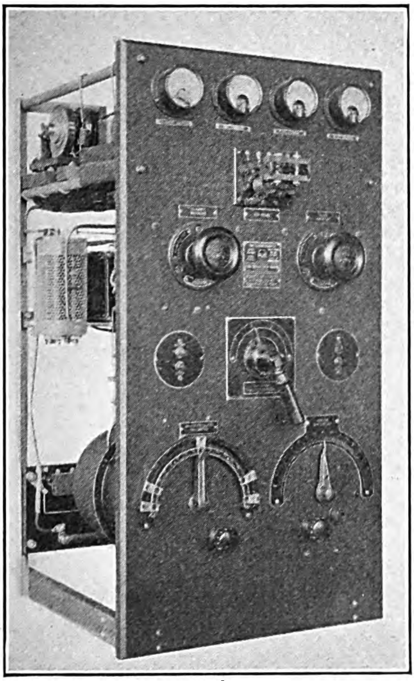

Fig. 7—Front View of Model ET-3627-A Transmitter.

L-1 which, as stated before, is provided with four taps and a variometer.

Since modern radio traffic conditions require rapid change-over from send to receive, a magnetically operated break-in relay is supplied in the transmitter. This relay is designated as *K*-2 in Fig. 6. It is equivalent to a double-pole, single-throw relay, with the parts designed to operate at keying speeds up to 40 words per minute. One pair of contacts are in series with the low side of the antenna circuit and serve to short circuit the input to the radio receiver during the transmitting condition. The second pair of contacts key the transmitter and are so timed that

they close slightly after and open slightly before the contacts in the antenna circuit. This arrangement prevents sparking at the antenna contacts and reduces the disturbance from clicks in the radio receiver.

The fundamental keying circuit may be understood by referring to the schematic diagram. The negative lead from the

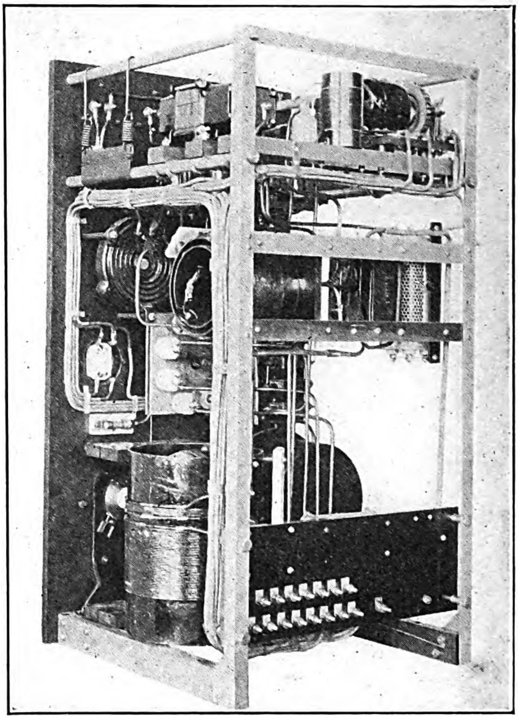

Fig. 8—Rear View of Model ET-3627-A Transmitter; Motor-Driven Chopper in Upper Right Section of Frame.

1000-volt generator connects to one of the key contacts on the break-in relay. After passing through these contacts, the negative plate circuit connects to the mid-point of the filament transformer *R*-2. In addition, the grid leaks *R*-1 and *R*-3 of the radio amplifier and master oscillator respectively are returned to the negative side of the plate circuit. This form of keying circuit therefore breaks both the negative plate circuit and the grid current with the result that a negative potential is impressed upon the grids of the tubes whenever the contacts open. Com-

note having constant characteristics is obtained with this modulating system and it has the additional advantage that it may be heterodyned at the receiving station if desired.

200-watt R.C.A. Transmitter

Fig. 7 shows a front view of the model ET-3627-A transmitter, developed for the Radio Corporation of America, which has been

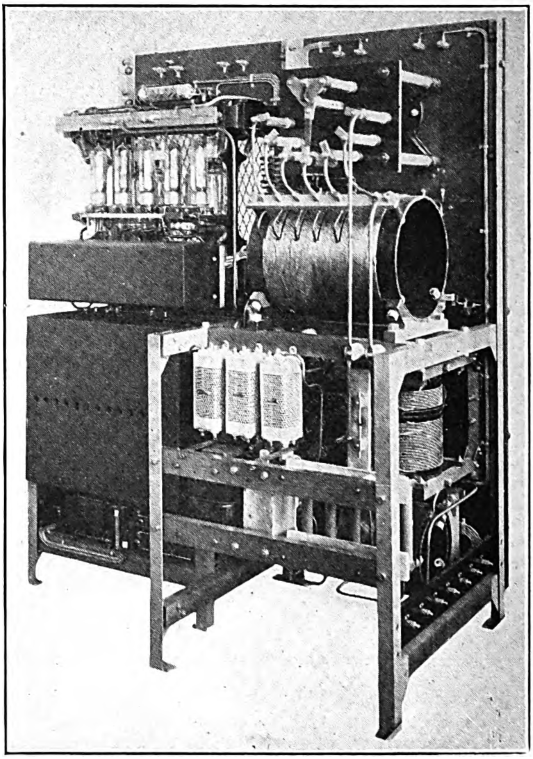

Fig. 10—Rear View of 500-watt Coast Guard Transmitter.

extensively applied for marine installations. The fundamental circuits in this transmitter are quite similar to those in the Coast Guard 200-watt equipments but some changes have been made to meet the requirements of commercial service. An adjustable positioning device for the master-oscillator variometer is of particular interest and is shown in the lower left section of the panel in Fig. 7. This device permits any five frequencies within the 312- to 500-kilocycle range to be selected and a

permanent adjustment maintained. Such an arrangement is of considerable help to an operator in changing from his calling to his working frequency as it makes it unnecessary to set the master-oscillator pointer carefully at an exact position on the dial.

ICW telegraph is carried out in the ET-3627-A transmitter

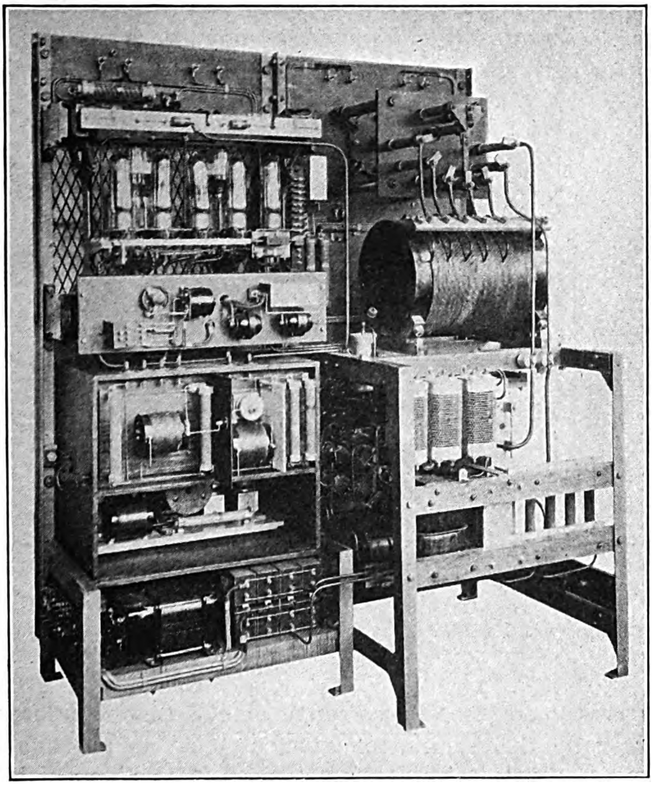

Fig. 11—Rear View of 500-watt Coast Guard Transmitter with Shielded Compartment Removed.

by means of a motor-driven chopper. Referring to Fig. 8, this chopper may be seen in the upper right section of the frame.

Extensive applications have been made of the ET-3627-A transmitter for coastwise vessels or for other ships which require a compact transmitter capable of giving a reasonable communication range. The daylight range under normal conditions from this transmitter when transmitting on CW is from 400 to 600 miles, while night ranges of 1500 miles have been obtained. These ranges are for transmission over water.

500-watt Coast Guard Transmitter

In order to meet the need for higher power equipment than provided for in the sets already described, a 500-watt transmitter was developed for the U. S. Coast Guard. A front view of this transmitter is shown in Fig. 9. Fig. 10 shows a rear view of the transmitter with the shielded compartment in position over the master-oscillator and audio-amplifier circuits, while Fig. 11 is a rear view with the shielded compartment removed.

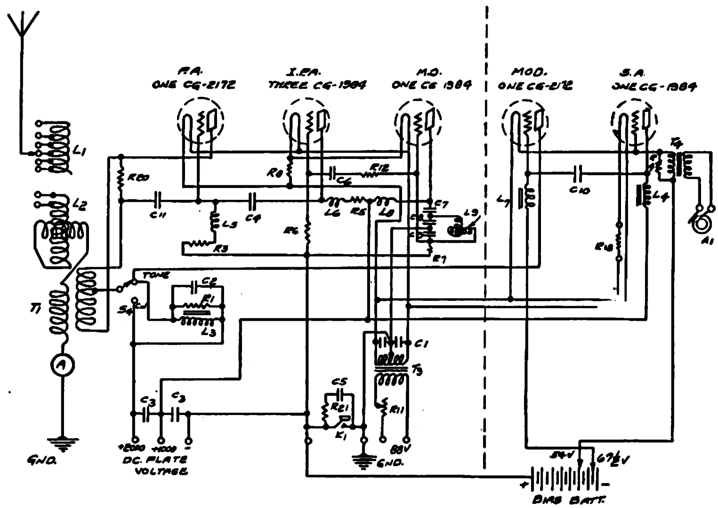

Fig. 12—Schematic Circuit Diagram Showing Fundamental Arrangement in 500-watt Transmitter.

A continuous frequency range of 125 to 500 kilocycles is provided by this equipment when used on an antenna whose capacitance falls within the limits of 0.00085 to 0.001 μfd. Since a transmitter of this power rating is ordinarily used on ships where fair-sized antennas may be erected, it is not designed to work into as low a value of capacitance as the smaller set. In addition, much lower transmission frequencies are possible. Signalling may be carried out by CW or ICW telegraphy, and the audio circuits in the transmitter are so designed that telephony may be carried on with slight modifications in the wiring.

The transmitter uses a total of seven vacuum tubes as follows:

1 CG-1984 (UV-211) as master oscillator.
3 CG-1984 (UV-211) as intermediate amplifiers.
1 CG-2172 (UV-851) as main radio amplifier.

1 CG-2172 (UV-851) as modulator.
1 CG-1984 (UV-211) as audio amplifier.

The functioning of these tubes will be described later with reference to the circuit diagram.

The CG-2172 tube is normally rated at 1 kw. output and operates on a d-c. plate voltage of 2000. The filament requires

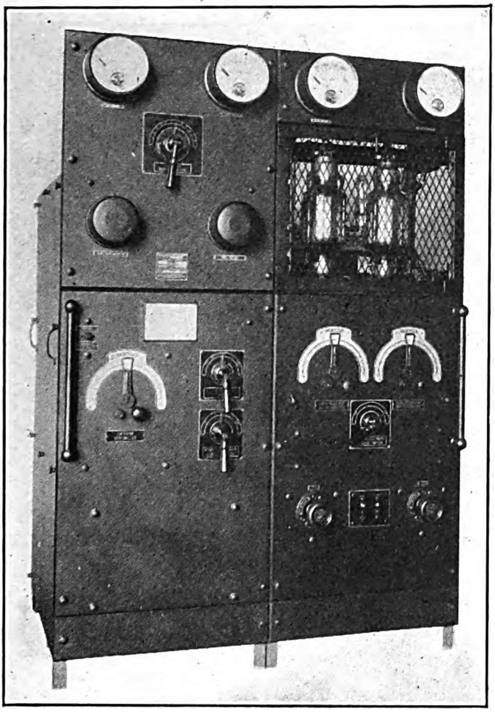

Fig. 13—View of 2-kw. Transmitter.

15.5 amperes at 11 volts and is heated from alternating current. The use of a 1 kw. tube in this transmitter to secure a nominal output of 500 watts was desirable for two reasons. In order to operate the antenna at the lower frequencies, a considerable amount of loading inductance is required with its corresponding losses. In order to provide ICW telegraphy, and at the same time permit telephony to be used later if desired, the CG-2172 tube was selected as a modulator. Actual tests on the completed transmitter showed that outputs of from 500 to 750 watts could easily be obtained even under unfavorable antenna conditions.

One of the chief electrical requirements for this transmitter was that it should permit any frequency within the specified band to be selected in a short period of time and that it should maintain this frequency constant to within 350 cycles despite normal changes in antenna characteristics or variations in line-supply voltage. Experience has shown that a high degree of frequency stability in a transmitter of this power can be

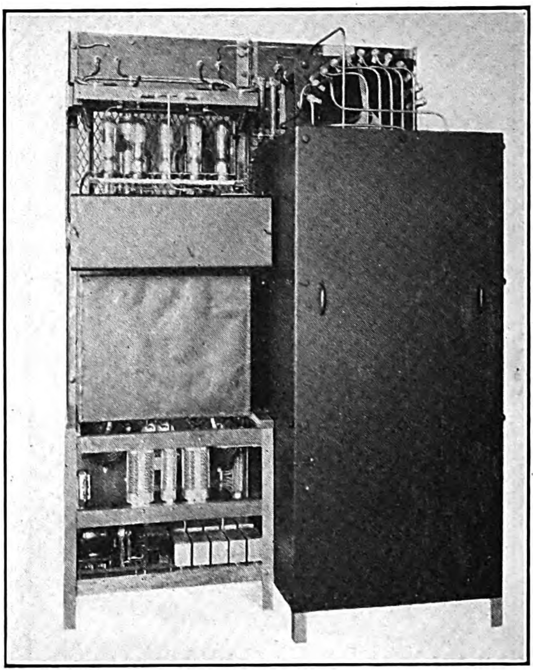

Fig. 14—View of 2-kw. Transmitter.

obtained through the use of an intermediate amplifier between the master oscillator and the main radio amplifier. This arrangement enables the master-oscillator or frequency-determining circuit to be built with relatively small well-shielded units, while the intermediate amplifier acted as a buffer to prevent reaction from the antenna circuit from disturbing the load on the master oscillator.

A schematic circuit diagram showing the fundamental arrangement in the 500-watt transmitter is shown in Fig. 12. The master-oscillator circuit is the capacitively-coupled type with inductance variation provided for changing the frequency. In

the transmitter proper two such variometers are utilized in order to spread out the frequency scale.

Referring to Fig. 12, the three intermediate power-amplifier tubes, which are connected in parallel, have in their plate circuits an inductor *L*-6 and a resistor *R*-5. The values of these two

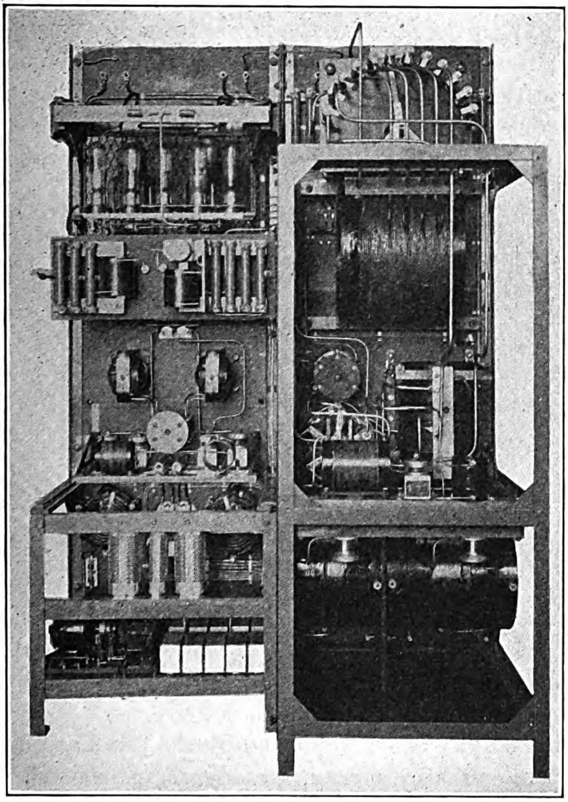

Fig. 15—View of 2-kw. Transmitter.

units are so selected that fairly uniform amplification is obtained over the entire frequency range of the transmitter. For this reason, no adjustments of any nature are required in the intermediate amplifier circuit. The voltage built up across *L*-6 and *R*-5 is used to excite the grids of the CG-2172 radio-amplifier tube, through the coupling capacitor *C*-4. A grid leak choke *L*-5 and a grid leak *R*-3 are provided in the grid circuit of the radio-amplifier tube. The plate circuit of the CG-2172 amplifier is coupled to the antenna circuit through the output transformer

T-1. In the actual transmitter, two of these transformers are supplied with an appropriate band change switch to place either transformer in circuit, depending upon the frequency desired. The characteristics of output transformers such as used in these transmitters are similar in some respects to standard power transformers. The primary or plate winding is designed with sufficient reactance so that the input to the power-amplifier tubes is limited until the antenna circuit is in resonance to the frequency being supplied to the grid of the amplifier. This provides simplified tuning for the radio operator as the power-amplifier tube does not draw its full load until the antenna circuit has been correctly

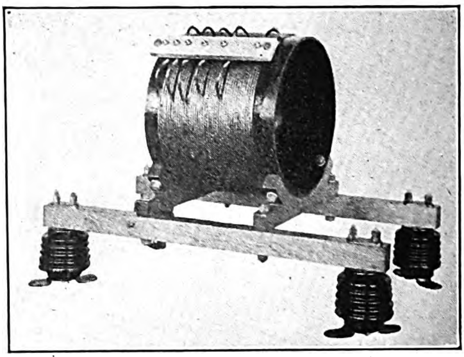

Fig. 16—External Loading Coil.

resonated. Loading of the antenna circuit is accomplished in the usual manner with a tapped inductor and a variometer for fine adjustment.

ICW telegraphy is accomplished by the familiar plate modulation system common to many broadcasting transmitters. The iron core reactor L-5, in Fig. 12, is common to the plate circuit of the radio-amplifier tubes and the modulator tubes. Audio frequency for modulating is obtained from a small tone alternator A-1 in the schematic diagram. The output from this machine passes through a step-up transformer T-4 where it is impressed upon the grid-filament circuit of the audio-amplifier tube. The audio-amplifier is of the reactance-capacitively-coupled type and delivers this output through capacitor C-10 to the modulator grid. In order to carry on telephony with this transmitter, it is merely necessary to substitute a microphone and battery for the tone alternator.

Referring again to Figs. 10 and 11, the general mechanical design of the 500-watt equipment may be observed. The master-oscillator, intermediate-amplifier, and audio-amplifier circuits are mounted in the left rear section of the transmitter. Complete shielding of these units is provided. The vacuum tubes are mounted above the audio-amplifier unit in a cradle which is cushioned with springs and sponge rubber. A door on the front

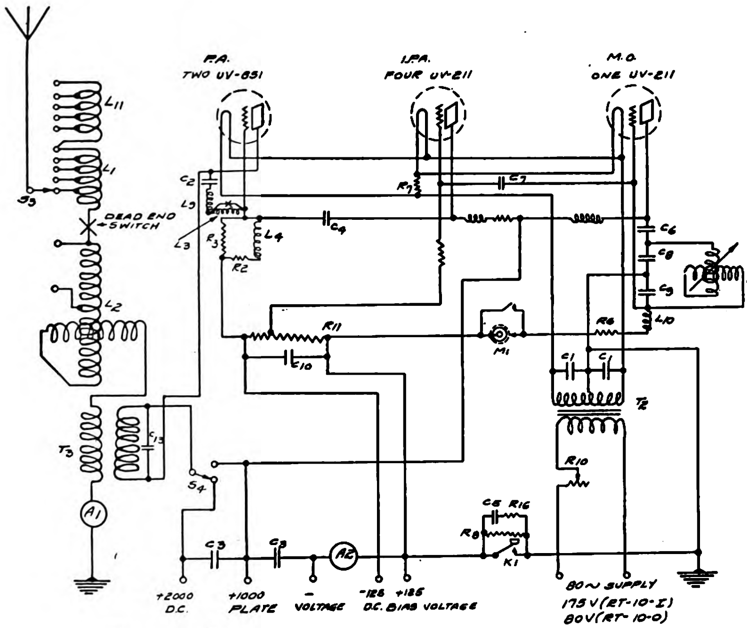

Fig. 17—Schematic Diagram of ET-3638 Transmitter.

of the panel provides for easy replacement of these tubes. The output transformers and the antenna-loading inductor with the antenna switch are mounted in the right rear section. Extensive use is made of Mycalex insulation on the antenna inductance switch. The band change switch which selects the appropriate output transformer is also insulated with Mycalex. As it is necessary to carry transmitters through narrow doors on shipboard, the design is such that easy disassembly of the transmitter into two or more sections can be accomplished. A small terminal board is built on one of the panels to provide for interconnection between units.

2-kw. R.C.A. Model ET-3638 Transmitter

For large vessels handling a considerable amount of traffic at long distances, a 2 kw. radio transmitter provides a logical rating. It is also suitable for shore installations, where comparatively long distance transmission is required. Views of a 2-kw. transmitter developed for the Radio Corporation are shown in Figs. 13, 14, and 15.

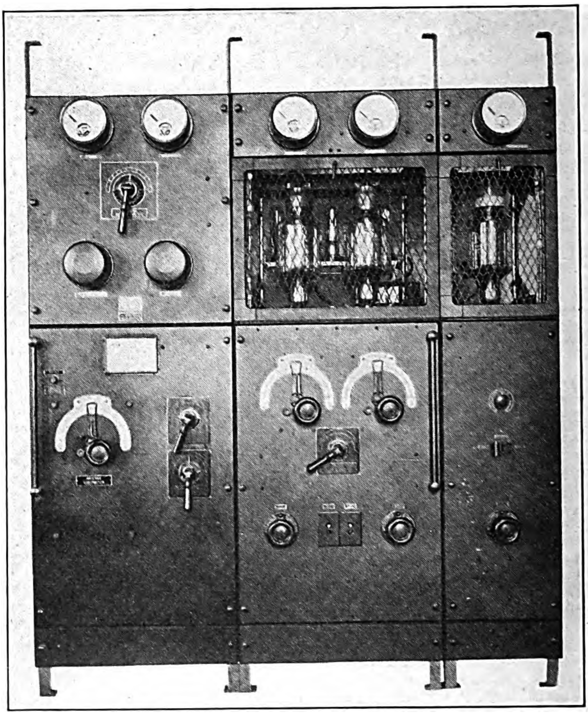

Fig. 18—Front View of 2-kw. Transmitter with Telephone Attachment.

The ET-3638 transmitter uses a total of seven vacuum tubes. Five of these tubes are type UV-211 and the remaining two are type UV-851. One of the UV-211 tubes operates as a master oscillator, four UV-211 function as intermediate amplifiers, and the two UV-851 tubes are connected in parallel as the main radio power amplifiers. A continuous frequency range of 125 to 500 kilocycles is covered by the transmitter when used on the average ship antenna. An antenna current of about 25 amperes is the usual value obtained. Continuous-wave and

interrupted continuous-wave telegraphy are provided, the latter being obtained by means of a motor-driven chopper.

Power supply for the transmitter is obtained from a 3-unit motor-generator set consisting of a motor with slip rings for filament heating, a 2000-volt 4.6 kw. plate generator, and a small bias generator for holding grid bias on the various tubes.

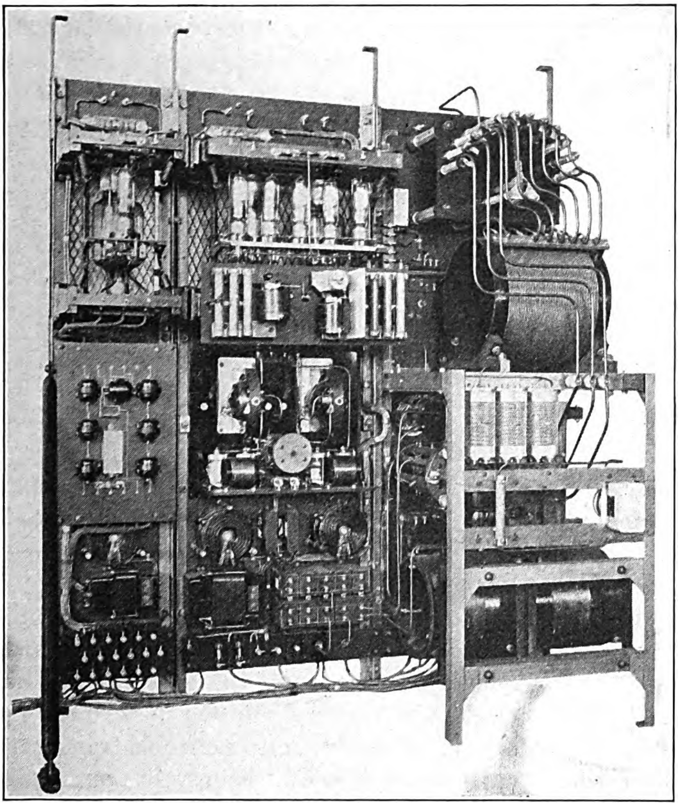

Fig. 19—Rear View of 2-kw. Transmitter with Telephone Attachment.

In order to permit operation at the lower radio frequencies on antennas whose capacitance is somewhat lower than the average, an external loading coil is supplied such as shown in Fig. 16. When operating into an 0.001 antenna at 125 kilocycles, the potential on the antenna end of the loading inductor reaches values as high as 32,000 volts. For this reason, ample insulation is provided on the loading inductors and on the antenna switch which changes the taps.

Extensive shielding has been employed in this 2 kw. transmitter not only to permit a high degree of frequency stability to be maintained, but also to enable the transmitter to be installed close to metal bulkheads without causing high losses. The various shields are easily removed for inspection or repair of the transmitter.

A schematic circuit diagram of the ET-3638 transmitter is shown in Fig. 17. This circuit is in most respects similar to those already described. The motor-driven chopper, for ICW, is used to break the grid leak current on the master oscillator tube. A cut-off bias obtained from the 125-volt bias generator is maintained on the grids of the four intermediate amplifiers and the two main amplifiers, so that the plate current of these tubes falls to zero whenever the excitation from the master-oscillator tube is interrupted by the chopper.

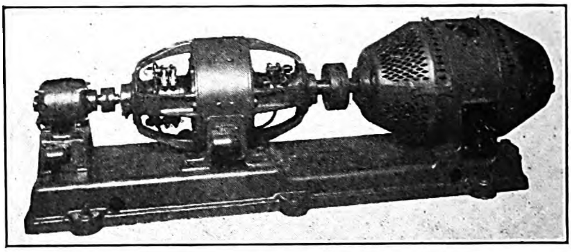

Fig. 20—Three-Unit Motor Generator Set.

A considerable amount of work has been done to minimize harmonic radiation from transmitters of this type. Suitable design of the output transformer and the antenna loading inductors, together with shielding, has resulted in very low radiation of harmonic energy. Measurements taken on a transmitter of this type show that the fundamental energy is approximately 30,000 times as great as the second harmonic and about 60,000 times as great as the third harmonic component. Referring to Fig. 17, the capacitor *C-13* in shunt to the primary of the output transformer *T-3* is used to by-pass the very high frequency harmonics.

U. S. Coast Guard Model T-2-A 2 kw. Transmitter

A 2-kw. transmitter with a telephone attachment was developed for service on some of the larger vessels of the U. S. Coast Guard. Front and rear view of this equipment are shown

in Figs. 18 and 19 respectively. The general arrangement of tubes in the radio transmitter proper is similar to that employed in the 500-watt Coast Guard equipment, with the difference that more tubes are used. The transmitter proper has one CG-1984 (UV-211) as a master oscillator, four similar tubes in parallel as intermediate amplifiers, and two CG-2172 (UV-851) tubes as the main radio amplifiers. In order to carry on telephony or ICW

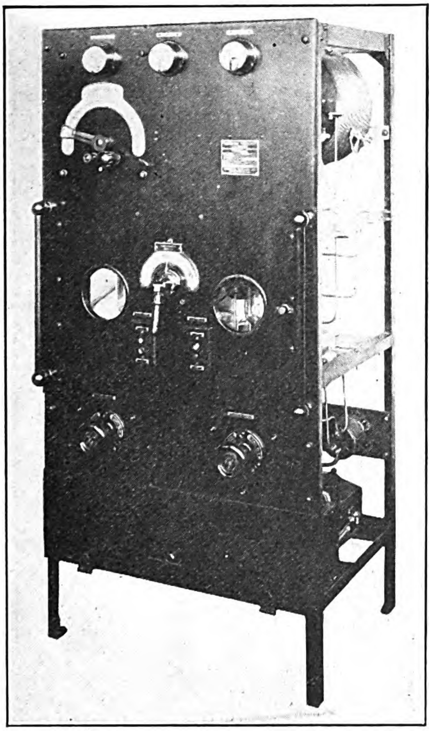

Fig. 21—400-watt Transmitter

telegraphy, a separate telephone attachment which may be mounted alongside the transmitter, is furnished. This attachment utilizes one CG-1984 as the speech amplifier and one CG-2172 tube as the modulator. When the signal switch on the transmitter is placed in the tone or ICW position, the plate voltage on all the tubes is reduced in order to permit the modulator tube to control properly the output of the radio amplifiers. The transmitter covers the same frequency range as the 500-watt Coast Guard set, namely, 125 to 500 kilocycles.

The type of 3-unit motor-generator set supplied with equipments of this nature is shown in Fig. 20. This machine is designed to carry its load when operating in high ambient temperatures such as are encountered in engine rooms.

Aircraft Radio Beacon Equipment

An interesting development being carried out at the present time consists of a system for guiding aircraft. It is known as the

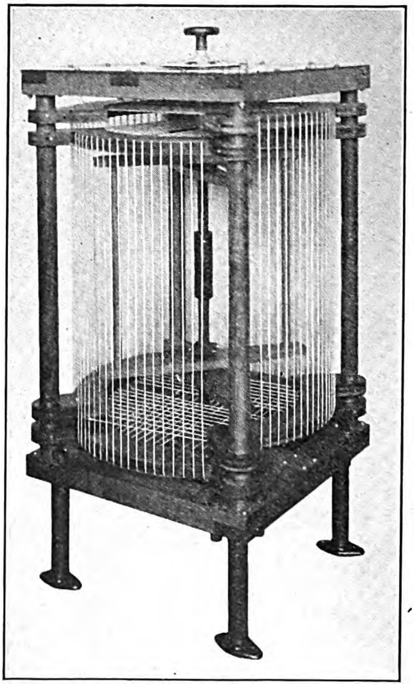

Fig. 22—Experimental Form of Goniometer.

double beam equi-signal method of transmission and permits the use of standard receiving apparatus in the aircraft. An additional feature is that the pilot or operator need not possess an extensive knowledge of the telegraphic code.

Two large loops are used for transmission instead of the conventional type of antenna. These loops are installed at right angles to one another and each produces, if separately energized, a well-known figure eight pattern of radiated energy. A goniometer, similar to the well-known Bellini-Tosi type, is utilized with

necessary modifications to adapt it best for this beacon system. The goniometer couples the radio transmitter to the transmitting loops and by variations of the rotor of the goniometer, the combined pattern of the loop may be rotated. In other words, the goniometer permits the combined pattern to be set for a desired course, with convenience from the transmitting station, without physically changing the position of the large transmitting loops. The rotor, or primary, of the goniometer consists of two coils which are alternately connected to the radio transmitter by means of an automatic relay. By suitable adjustment of the angle between the

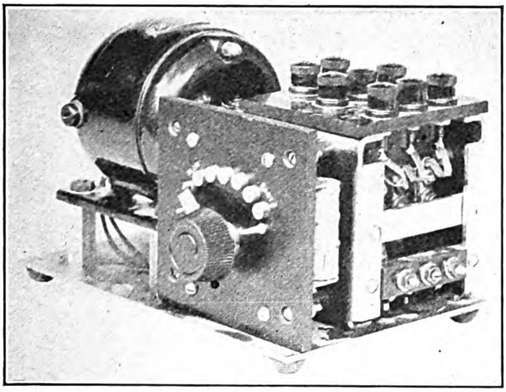

Fig. 23—Automatic Signalling Device.

two rotor coils, the patterns of the two loops are made to overlap, and in this overlapping zone what is known as an equi-signal area is maintained.

In order to produce interlocking signals, the automatic relay may be arranged to send for example, the letter *N* on one rotor coil and the letter *A* on the other coil. By suitable timing, the dot of the *A* is made to begin as the dash of the *N* terminates, with the result that the receiving operator hears a series of dashes, as long as he is flying in the equi-signal zone. Tests have shown that the width of this equi-signal zone is to the order of 2 to 3 miles at distances of about 100 miles from the transmitting station and this width can be varied by suitable adjustment at the goniometer.

Fig. 21 shows a 400-watt transmitter which is being used for some of the experimental work with the aircraft beacon. This transmitter utilizes two UV-204-A tubes in a self-rectified circuit with 500 cycles plate supply and is adjusted to operate on a

frequency to the order of 290 kilocycles. An experimental form of the goniometer is shown in Fig. 22. The two inside or rotor coils may be clearly seen, while the secondary or outer coils which consist of several turns are wound at right angles to one another and connect to each of the transmitting loops. The automatic signalling device shown in Fig. 23 is provided with cams for

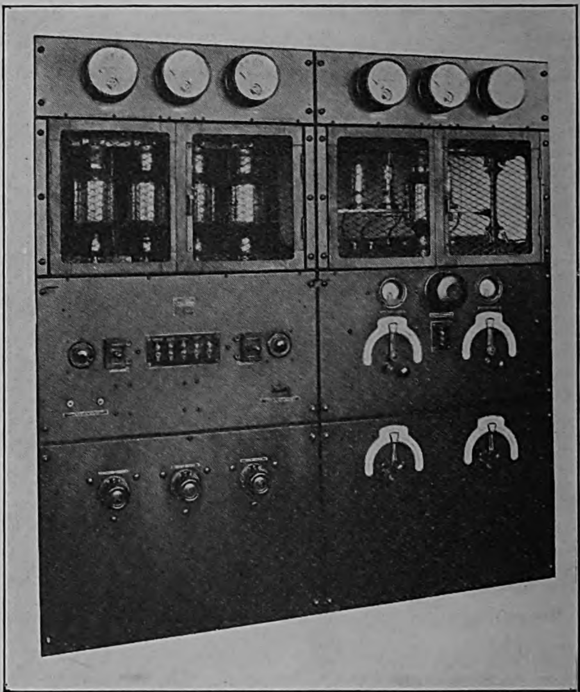

Fig. 24—View of 1-kw. Broadcast Transmitter.

sending the appropriate letters to secure the interlocking signal. A small rheostat is provided to vary the speed of the motor which drives this signalling device.

An installation of the radio beacon equipment has been made at Hadley Flying Field, New Brunswick, N. J., and with the aid of this equipment, practical operating data and the general characteristics of the beacon system are being obtained. This work is being carried on in cooperation with the Department of Commerce and active steps are being taken to work out a practical system that may be used throughout the country.

Various views of the 1-kw. broadcast transmitter are shown in Figs. 24, 25, and 26. The condenser microphone with its self-contained amplifier is shown in Fig. 27.

The transmitter as normally designed covers a frequency range from 666 to 1200 kilocycles, although it may be easily

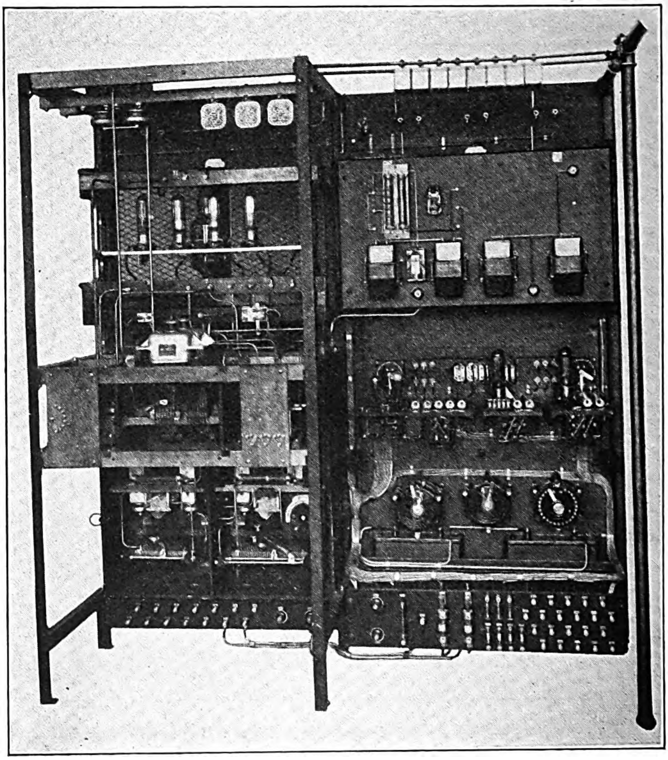

Fig. 26—View of 1-kw. Broadcast Transmitter.

modified for any frequency within the broadcast band. The complete equipment utilizes the following vacuum tubes:

2 UX-201-A as audio amplifier.
1 UX-210 as audio amplifier.
2 UV-211 as audio amplifiers.
4 UX-851 as modulators.
1 UV-211 as master oscillator.
2 UV-211 as intermediate radio amplifiers.
1 UV-851 as the main radio amplifier.
1 UV-211 as oscilloscope rectifier.

All of the above tubes are contained within the transmitter proper with the exception of one of the UX-201-A tubes which is mounted with the condenser microphone.

It will be observed that four modulator tubes are used and one main radio-amplifier tube. This ratio permits maximum modulation to be obtained without overshooting the modulator tubes. Due to the use of such a low-impedance bank of modulator tubes,

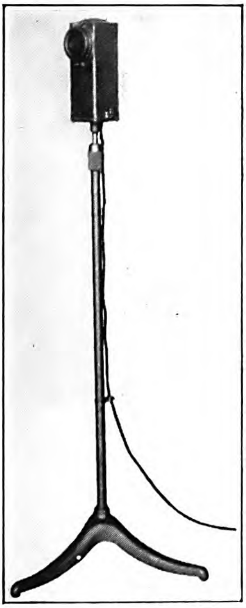

Fig. 27—Condenser Microphone with Self-Contained Amplifier.

combined with a large modulation reactor, the low audio frequencies are well maintained.

A schematic circuit diagram is shown in Fig. 29. All the vacuum tubes to the left of the antenna circuit in this diagram are in the audio-frequency circuit, while the tubes to the right are in the radio-frequency circuit. The condenser microphone works into the first UX-201-A tube, this tube being mounted with its associated output transformer, in the microphone housing. The secondary winding on the output transformer is of low impedance so that the microphone cable may be run for a considerable distance without picking up interfering current. The first three stages of audio amplification in the transmitter proper are the resistance type, and the fourth stage utilizes an iron core

reactor in its plate circuit. Volume-control potentiometers are provided on the input to the third and fifth stages of amplification. Provision is also made for ready connection to incoming lines for outside pick-up service.

Referring to the radio-frequency tubes in the schematic diagram, the master oscillator has its frequency controlled by means of a variable capacitor for fine adjustment and by variation of taps on the inductor for coarse adjustment. This tube supplies grid excitation to 2 UV-211 tubes which are provided with a tuned tank circuit and function as intermediate amplifiers. The main radio-amplifier tube, a UV-851, also has its plate circuit tuned and is in turn inductively coupled to the antenna

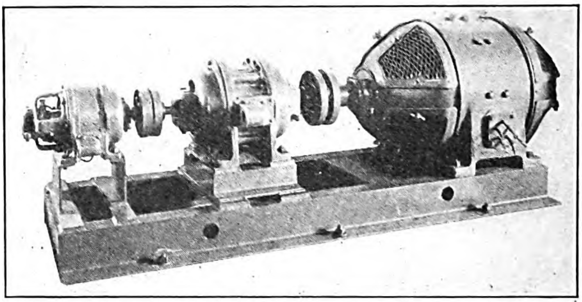

Fig. 28—Three-Unit Plate Motor Generator Set.

circuit. Complete shielding is employed to provide maximum frequency stability and the various coupled circuits minimize harmonic radiation.

One of the problems in the successful operation of a broadcast transmitter consists of reducing all hum or background of noise in the carrier wave to an extremely low value. In the ET-3633 equipment, this is accomplished by the use of suitable filter for the plate circuit of the vacuum tubes and by the use of a separate direct-current filament generator for all the larger tubes. The audio amplifiers have their filaments heated from a storage battery. Fig. 28 shows the three-unit plate motor generator set and Fig. 31 the filament motor generator set.

The oscilloscope supplied with the equipment for checking the percentage modulation is of interest. This unit is shown in Fig. 30 and is equivalent to the familiar oscillograph with the exception that it contains but one vibrator and is designed to use an incandescent lamp as a light source. The UV-211 tube which is con-

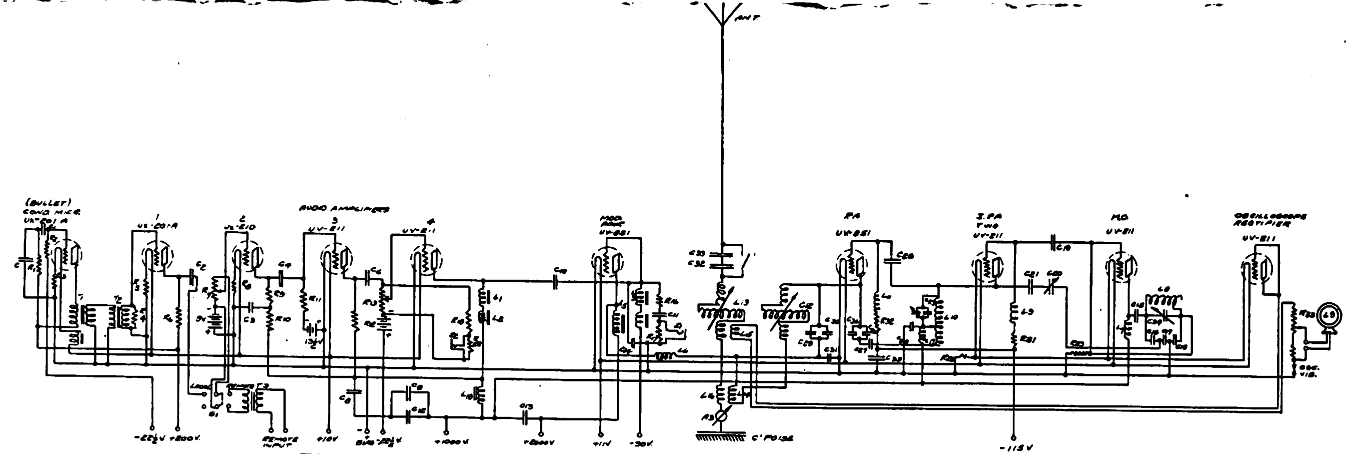

Fig. 29—Schematic Circuit Diagram of 1-kw. Broadcast Transmitter.

A_3 Ant. Ammeter
C Cond. Microphone
C_1 Grid Cond.
C_2 Coupling Cond.
C_3 By-Pass Cond.
C_4 Coupling Cond.
C_6 Coupling Cond.
C_8 Filter Cond.
C_{10} Coupling Cond.
C_{11} Blocking Cond.
C_{12} Filter Cond.
C_{13} Filter Cond.
C_{14} By-Pass Cond.
C_{15} M.O. Plate Blocking Cond.
C_{16} M.O. Plate Cond.
C_{17} M.O. Plate Cond.
C_{18} M.O. Grid Cond.
C_{19} I. P. A. Coupling Cond.
C_{20} Neutralizing Cond.
C_{21} Neutralizing Blocking Cond.
C_{22} I.P.A. Var. Tank Cond.
C_{23} I.P.A. Tank Cond.
C_{24} I.P.A. Plate By-Pass Cond.
C_{25} P.A. Coupling Cond.
C_{26} P.A. Neut. Cond.
C_{27} P.A. Neut. Blocking Cond.
C_{29} P.A. Tank Cond.
C_{30} P.A. Tank Cond.
C_{31} P.A. Plate By-Pass Cond.
C_{32} Ant. Series Cond.
C_{33} Ant. Series Cond.
C_{34} M.O. Var. Tank Cond.
C_{35} Bias Filter Cond.
C_{36} P.A. Neut. Cond.
L_1 Plate Reactor
L_2 Plate Reactor
L_3 Mod. Grid Reactor
L_4 Mod. Grid Reactor
L_5 Modulation Reactor
L_6 P.A. Plate Choke
L_7 M.O. Plate Choke
L_8 M.O. Tank Induct.
L_9 I.P.A. Grid Choke
L_{10} I.P.A. Tank Induct.
L_{11} P.A. Grid Choke
L_{12} P.A. Tank Var.
L_{13} Ant. Variometer
L_{14} P.A. Coupling Induct.
L_{15} Oscill. Pick-Up Induct.
L_{16} Ant. Coupling Induct.
L_{17} I.P.A. Plate Choke
R_1 Polarizing Res.
R_2 Grid Res.
R_3 Fil. Res.
R_4 Transformer Loading Res.
R_5 Fil. Res.
R_6 Coupling Res.
R_7 Volume Control Res.
R_8 Fil. Res.
R_9 Plate Res.
R_{10} Plate Series Res.
R_{11} Grid Res.
R_{12} Plate Res.
R_{13} Volume Control Res.
R_{16} Listening Res.
R_{17} Listening Res.
R_{18} Listening Res.
R_{19} Parasitic Res.
R_{20} Parasitic Res.
R_{21} Parasitic Res.
R_{22} Listening Res.
R_{23} Parasitic Res.
R_{24} Parasitic Res.
R_{25} Parasitic Res.
R_{27} Parasitic Res.
R_{28} Filament Res.
R_{29} M.O. Grid Leak Res.
R_{31} I.P.A. Grid Leak Res.
R_{32} P.A. Grid Leak Res.
R_{33} Oscill. Resis.
S_1 Microph. Switch
T_1 Bullet Output Transformer
T_2 Input Transformer
T_3 Input Transformer
J_1 Listening Jack
J_2 Listening Jack
L.S. Loud Speaker.

nected as a rectifier in the transmitter is coupled to the antenna circuit and is utilized to supply the audio frequency to the oscilloscope.

The first transmitter of the ET-3633 type is in service at station CYJ in Mexico City. Similar transmitting equipment

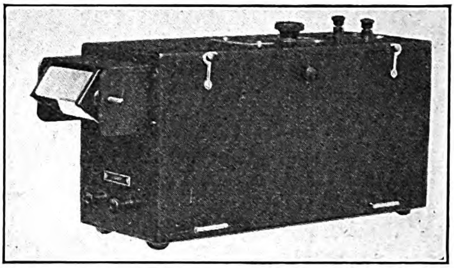

Fig. 30—Oscilloscope Supplied with Equipment for Checking Percentage Modulation.

has been built for Cornell University and St. Lawrence University.

Crystal-Controlled Broadcast Amplifier

A 1-kw. crystal-controlled amplifier which is used in some of the General Electric broadcast stations is shown in Figs. 32 and 33. This unit supplies grid excitation to the water-cooled high-

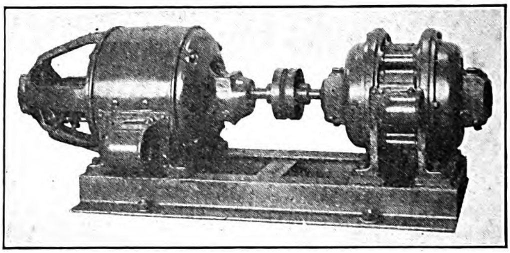

Fig. 31—Filament Motor Generator Set.

power radio-frequency amplifiers. A schematic diagram is shown in Fig. 34 and illustrates the various circuits which are employed in the crystal-controlled unit. The crystal-controlled tube, type UX-210, operates with the crystal connected between its grid and filament circuits. The plate circuit of the tube is tuned by means of a variable condenser which is designed to cover the broadcast-frequency band. The crystals themselves are mounted in a temperature-controlled compartment and a thermostat is

supplied in order to maintain the temperature constant at 45 deg. C. Provision is made for mounting four crystals, any one of which may readily be selected by means of a switch on the panel.

A second UX-210 tube is used to amplify the output from the crystal-controlled tube and this in turn is followed by a UV-211, 50-watt tube. Two additional UV-211 tubes connected in

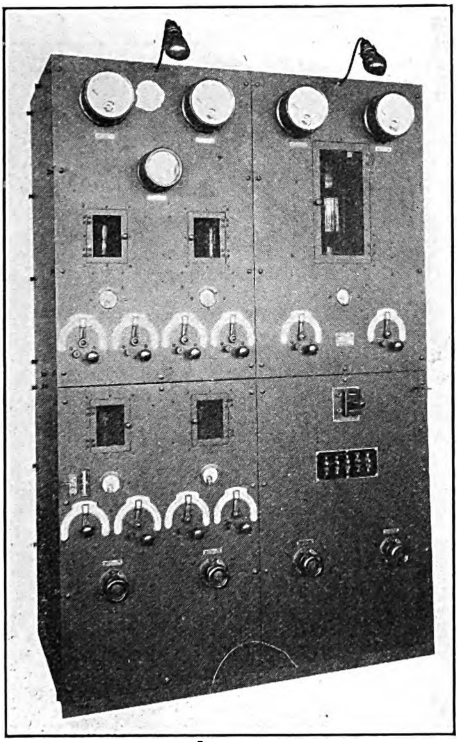

Fig. 32—Front View of 1-kw. Crystal-Controlled Amplifier.

parallel amplify the output from the first UV-211 tube and these are followed by a 1 kw. tube type UV-851. Straight amplification is employed throughout this unit, the crystal being ground for the final output frequency that is desired. Sufficient energy is available from the UV-851 stage to excite 1 or 2 water-cooled radio-amplifier tubes.

1-kw. High-Frequency Transmitter

In order to investigate the transmission characteristics of some of the higher frequencies, a 1 kw. transmitter was developed

to cover a continuous frequency range from 3748 to 14990 kilocycles (80 to 20 meters). This transmitter was designed so that crystal control could be employed for any desired frequency within the range, and in addition master-oscillator control was supplied so that gradual and continuous variation of frequency was possible. A number of plug-in coils were utilized so that efficient operation throughout the band could be maintained.

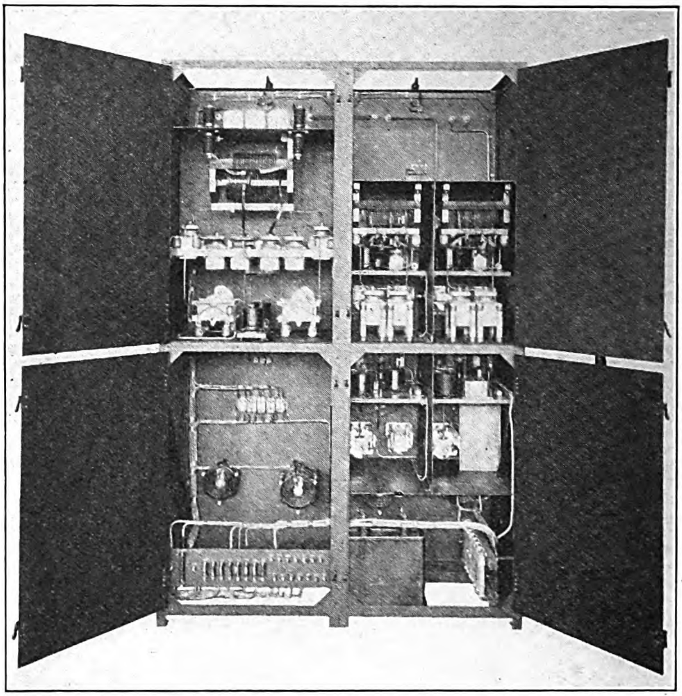

Fig. 33—Rear View of 1-kw. Crystal-Controlled Amplifier.

A front view of the 1 kw. transmitter is shown in Fig. 35. The view in Fig. 36 is taken with the shielding door open, which provides access to all the removable coils and also the vacuum tubes. Each amplifier stage is enclosed in its own shielded compartment in order to secure satisfactory stability.

The various circuits in this high-frequency transmitter are covered in schematic form in Fig. 37. A total of six vacuum tubes are used. The first or crystal-controlled tube is known as type SA-14 and is similar to the standard UX-210 except that it has a higher amplification constant and higher plate impedance.

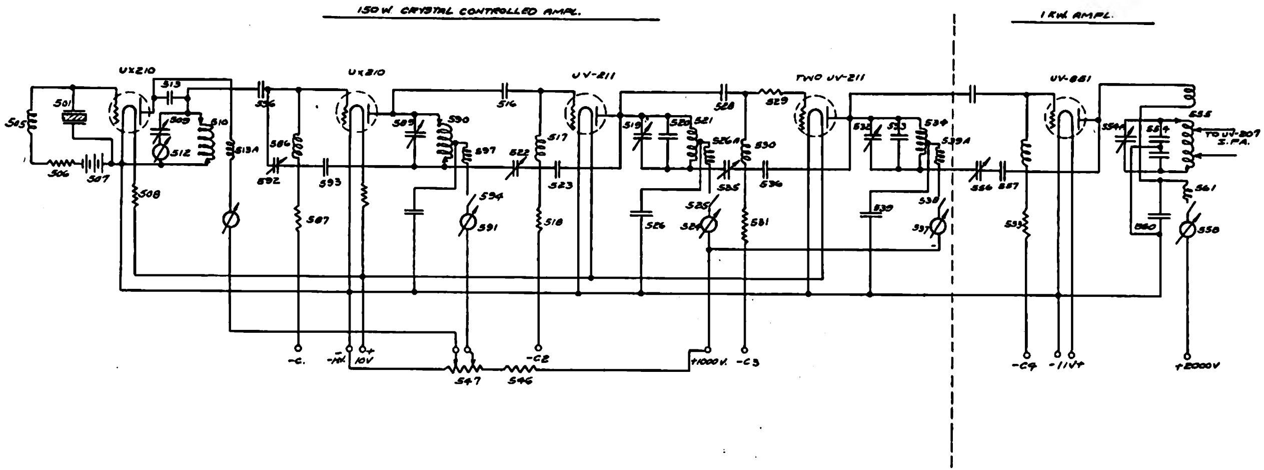

Fig. 34—Schematic Diagram Illustrating Various Circuits Employed in Crystal-Controlled Unit.

The crystal-control tube excites a second SA-14 tube, the plate circuit of this tube being tuned to some multiple of the crystal frequency. In some cases the third harmonic of the crystal frequency is utilized while for the higher frequencies the fifth harmonic is selected. The third tube in the transmitter is a standard UX-210, and is designed either to amplify the output from the preceding stage in the case of crystal control or to act as a self-

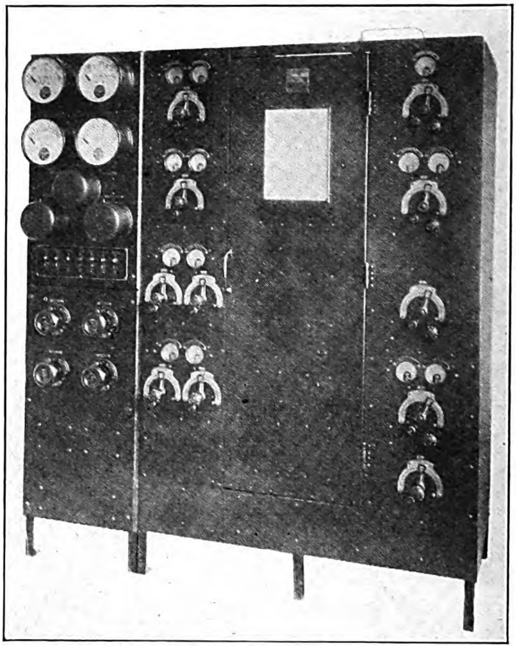

Fig. 35—Front view of 1-kw. Transmitter.

excited master oscillator when crystal control is not desired. A double-pole, double-throw switch provides for quick changeover from crystal to master operation or vice versa.

The remaining three stages consist of a UV-211, a UV-204A, and a UV-851 tube respectively. Each of these stages is designed with a tuned plate circuit inductively coupled to the following grid circuit. Variable condensers are provided so that the various stages may be balanced to prevent self-oscillation.

Factors Influencing the Type of Circuit

In any discussion of the various types of transmitting equipment, such as described in this paper, the question may be asked

what factors chiefly influence the selection of one type of output circuit as compared to another, when intermediate stages of amplification are necessary, etc. It has been our experience that for commercial equipment which is required to cover a fairly wide frequency band with a minimum number of controls, an antenna transformer, such as shown in Fig. 6 in the circuit diagram, is preferable. With a properly-designed antenna trans-

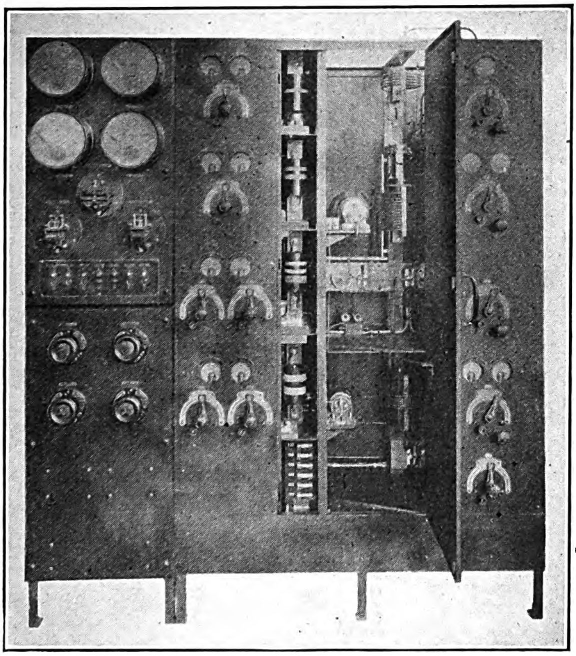

Fig. 36—View of 1-kw. Transmitter with Shielding Door Open.

former, the tubes are better protected against overload in case the circuits are not correctly resonated and frequency ranges having a ratio of approximately 2 to 1 may be covered without any adjustments whatever being made on the antenna transformer.

In the case of very high frequency transmitters where distributed capacitance of the circuit becomes increasingly important, it seems desirable to tune definitely the plate circuit of the final amplifier tube in order to secure best operation. We then have a so-called *tank* output circuit instead of the untuned antenna transformer arrangement.

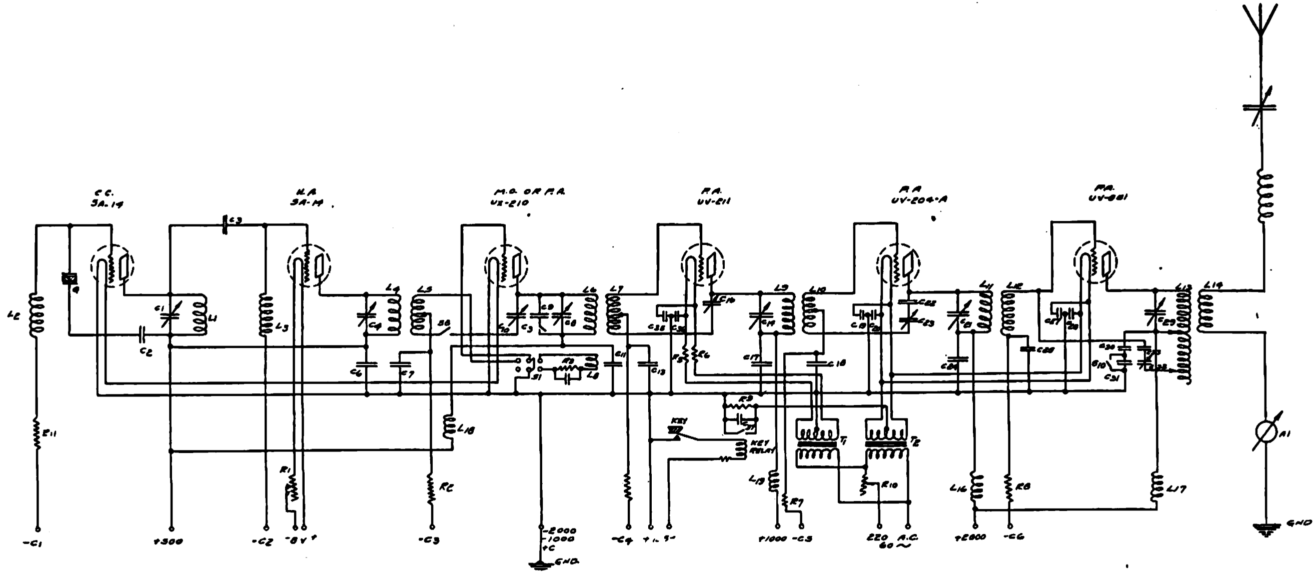

Fig. 37—Various Circuits of 1-kw. High Frequency Transmitter.

A_1 Ant. Ammeter
C_1 Var. Tank Cond.
C_2 By-Pass Cond.
C_3 Coupling Cond.
C_4 Var. Tank Cond.
C_6 By-Pass Cond.
C_7 By-Pass Cond.
C_8 Var. Tank Cond.
C_9 Tank Cond.
C_{10} Neutralizing Cond.
C_{11} By-Pass Cond.
C_{12} M. O. Grid. Cond.
C_{13} By-Pass Cond.
C_{14} Var. Tank Cond.
C_{16} Var. Neutralizing Cond.
C_{17} By-Pass Cond.
C_{18} By-Pass Cond.
C_{19} Fil. By-Pass Cond.
C_{20} Fil. By-Pass Cond.
C_{21} Var. Tank Cond.
C_{22} Blocking Cond.
C_{23} Var. Neutralizing Cond.
C_{24} By-Pass Cond.
C_{25} By-Pass Cond.
C_{27} Fil. By-Pass Cond.
C_{28} Fil. By-Pass Cond.
C_{29} Var. Tank Cond.
C_{30} By-Pass Cond,
C_{31} By-Pass Cond.
C_{32} Var. Neutralizing Cond.
C_{33} Blocking Cond.
C_{34} Ant. Series Cond.
C_{35} Fil. By-Pass Cond.
C_{36} Fil. By-Pass Cond.
C_{37} Key Cond.
L_1 Tank Induct.
L_2 Grid Choke—C.C.
L_3 Grid Choke—H.A.
L_4 Tank Induct.
L_5 Grid Induct.
L_6 Tank Induct.
L_7 Grid Induct.
L_8 M. O. Grid Induct.
L_9 Tank Induct.
L_{10} Grid Induct.
L_{11} Tank Induct.
L_{12} Grid Induct.
L_{13} Tank Induct.
L_{14} Ant. Coupling Induct.
L_{15} Ant. Loading Induct.
L_{16} Plate Choke
L_{17} Plate Choke
L_{18} Plate Choke
L_{19} Plate Choke
Q Mounted Quartz Crystal
R_1 Fil. Resistance
R_2 Grid Leak Res.
R_3 Grid Leak Res.
R_4 Grid Leak Res.
R_5 Fil. Resistance
R_6 Fil. Resistance
R_7 Grid Leak Resis.
R_8 Grid Leak Resis.
R_9 Key Resistance
R_{10} Fil. Control Res.
T_1 Fil. Transformer
T_2 Fil. Transformer

The power rating of a transmitter determines to a large extent whether or not intermediate stages of amplification are required. In order to secure a high degree of frequency stability in a master-oscillator, power-amplifier type of set, it is essential that considerable circulating energy be maintained in the master circuit. In addition, it is desirable to make the capacitance and inductance elements in the master circuit so that they maintain their values over long periods and so that they may be easily shielded. This ordinarily results in a fairly small master-oscillator assembly with a 50-watt tube as the usual type. If a 50-watt tube is too small to provide sufficient excitation for the main power amplifier in a transmitter, it is then desirable to introduce an intermediate stage of amplification or perhaps several stages. In the 2 kw. type of transmitter such as previously described, the intermediate amplifier is so designed that it functions without adjustment over the complete frequency range of the transmitter.

In the case of crystal-controlled transmitters, cascade-amplifier circuits are necessary due to the low level obtained from the crystal-controlled tubes. For high-frequency transmitters where, of course, crystals are chiefly used, developments are now under way with transmitters which will be equipped with special high-frequency tubes. Such tubes similar to the new UV-852 tubes will be designed to have low inter-electrode capacities which make them more suitable for operation at the higher frequencies than the present standard tubes.

The problem of measuring the output of high-frequency transmitters by means of a dummy load has been given considerable thought. If an attempt is made to use the conventional dummy antenna resistors, it is found that their inductance has an appreciable effect on the load circuit, and accurate measurements are difficult. One method which has been successfully used to measure high-frequency power consists of a bank of incandescent tungsten lamps which are used to load the transmitter. These lamps are mounted in a compartment with a photo-electric cell and their brilliancy controls the internal resistance of the cell. Such a device may be calibrated on direct current or low-frequency alternating current and tests have shown that accurate measurements can be made with such a system. In other words, the brilliancy of the lamp when heated by high frequency is the same as when heated by an equivalent power at low frequency or by direct current.

APPARENT NIGHT VARIATIONS WITH CROSSED-COIL RADIO BEACONS*

By

Haraden Pratt

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Summary—*The combined effects of apparent wave direction shifts and fading, of signals from a crossed coil type of radio beacon as received on airplanes in flight at night, are described. A brief explanation of the operation of such a beacon is given. The results of observations of similar signals at night received on an automobile, together with some general conclusions, are mentioned.*

It early became apparent that the employment of coil antennas on airplanes for determining the directions or bearings of radio stations was beset with difficulties. The coils were necessarily limited in size and considerable amplification was needed to secure a signal of suitable strength. Airplane engine ignition interferences in the radio receiving system prevented the use of very much amplification and so the distance range was limited. The high level of noise present on aircraft prevented a close observation of the minimum signal sector during rotation of the coil, thereby impairing the accuracy of that method as applied to aeronautical navigation; not to mention many other disqualifying practical and operational difficulties.

While the complicated system of taking several bearings on airplane radio signals from ground stations and communicating a position back to the airplane has found successful application in Europe, various means for the taking of radio station bearings from airplanes were attempted without reaching a practical solution of this problem. It was not, however, until the crossed coil radio beacon was proposed that those interested in the art became encouraged to expect the realization of a simple and workable system to guide aircraft.

The crossed coil beacon idea was investigated by the Bureau of Standards several years ago, and further work on it carried forward by the Signal Corps, since which time its development has been rapid, and there are now several operating beacons installed in the eastern part of the United States.

* Original Manuscript Received by the Institute, March 24, 1928. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce.

Essentially, this beacon comprises two coil antennas disposed in two vertical planes fixed at an angle to each other. In a simple form of the beacon, this two-coil system is free to be rotated about a vertical axis. When the coils are similarly excited with radio-frequency currents modulated at an audio rate, signals of equal intensity from each will be heard on a receiving set when situated along any one of the two vertical planes bisecting the angles between the planes of the coils. At other points the signal intensities from each coil will be different. This marking off in space of vertical equisignal planes constitutes the directive feature of the crossed coil type of beacon, and these equisignal zones are frequently referred to as the courses set down by it. Ordinarily, only one of these courses is used.

In its practical application and for convenience in construction the beacon used in this country employs a radiating system of two large fixed single-turn loops disposed at right angles to each other. It is possible to rotate the radio field about these antennas without rotating the antennas. A suitable goniometer interposed between these loops and the power source permits, by the turning of its rotor, the rotation in space of the equisignal zones. A mechanical device automatically transmits the letter *A* on one coil and the letter *N* on the other. These are so interlocked that a continuous buzz or dash is heard along the equisignal zone. The width of this zone where the perfect continuous dash only is heard depends largely upon the acuteness of the observer's attention, and may vary between limits, for example, of from $1\frac{1}{2}$ to 3 degrees. To the right or left of this zone either the letter *A* or *N* predominates distinctly.

For the guidance of aircraft the crossed coil beacon possesses several obvious advantages:

(1) There is no zone of minimum or maximum signal strength to be observed.

(2) Location of the course or beam is secured by an automatic comparison of two signals.

(3) Regardless of the position of the aircraft within a wide angle when off the course this beacon furnishes a definite signal enabling the craft to locate and return to its proper course. This characteristic permits temporary detours to be made during flight around stormy areas or obstructions, a very important and valuable feature.

(4) The aircraft uses an ordinary receiving set with the usual

trailing wire antenna permitting signals to be received under the most favorable conditions as regards minimum sensitivity of airplane receiving set to local noise and disturbances.

(5) An aircraft may be guided along a set and invariable airway without reckoning wind drift.

These advantages had much to do with the selection of this type of beacon for application to our national airway systems particularly those over which schedule mail airplanes operate.

To secure information of a practical nature, receiving equipment was installed in the summer of 1927 on a mail airplane operating between the airports of Cleveland and New York over an airway approximately 380 miles long. Two directive radio beacons were available, one at the New York terminal and the other at a point 170 miles west near Bellefonte, Pa. The radio frequency used was 290 kc. The audio modulation was at the rate of 500 cycles per second. The radiating loops were of a triangular shape 300 feet long and 80 feet high at the apex. The current in each loop was 8 amperes. This route crosses the Allegheny mountain ranges which occupy nearly all of that region and is a particularly favorable one for this experimental flying both because of the rough topography and the prevalent foggy and cloudy weather.

The accuracy of the directive beacon as a guide had been established as of a high order, through considerable use in the past, but upon examination of the situation it was learned that all previous experience with it had been confined to daylight flying. Its operation at night had been untried and it was not known to what extent the well-known night shift phenomenon previously observed with direction-finding systems might affect it. Night flights over the airway mentioned were therefore undertaken in August, 1927, and the writer immediately observed results inconsistent with those secured by day. During the first flight it seemed impossible to keep the airplane on a course corresponding to the interlocking dash signal. No sooner was this signal received when it would change to the letter *A* or letter *N*. No amount of manipulation of the airplane would improve the situation. The dash signal would come and go at intervals of a few minutes. It soon became apparent that at the distance involved, which was about ninety miles, no accurate fixed course existed, but that the equisignal zone was rapidly moving about in an indefinite way.

This disconcerting effect indicated that some further study of the phenomenon must be made, and several night flights up to distances of 175 miles from the beacon were undertaken. In every case the shifting of the equisignal zone or course was noticed. The general results secured from five flights made at an average altitude of 2000 feet may be stated as follows:

(1) Within 25 miles of the beacon the shifting was not of a very serious nature.

(2) At 50 miles the shifting became pronounced but due to the zone appearing to be stationary in its proper position for possibly 75 per cent of the time, the beacon could still be depended upon when used with judgment.

(3) At a distance of 100 miles the shifting became very pronounced and persisted for more than 50 percent of the time, giving the beacon a questionable value.

(4) At 125 miles the beacon was of no further use as a guide.

(5) The shifting of the zone was gradual so that at first one would be inclined to think it due to the movement of the airplane.

(6) It appeared that the topography of the country between the beacon and the airplane exerted a considerable influence on the extent of the variation.

(7) Exceptional variations in shift over an arc as great as 100 degrees in azimuth were noted, but in general the change was confined to within possibly 25 degrees.

(8) Beyond 15 miles the fading of the general level of signal received was very severe during flight over mountains. Several variables being involved, no conclusions have been reached as to the relative contribution of each factor.

A few observations of the Bellefonte beacon at night have been made on the ground at Washington, a distance of 134 miles. While present, the shifting phenomenon was less pronounced than that observed in the air. On a night flight from Harrisburg to Washington no shifting of the zone was noticed, using signals from a beacon at College Park, Md. As there are no marked mountain ranges near College Park, these observations would indicate that topographical features have an important bearing on the matter of these variations.

To shed further light on the question, two sets of night-time measurements on the Bellefonte beacon were made in October 1927 by an automobile party, one at a point 22 miles and another

32 miles distant, both locations being in mountainous territory. Records obtained with a graphic field intensity recorder were made of signals transmitted by one Bellefonte loop, using both a vertical antenna and a coil antenna placed in a vertical place extending towards the beacon. Ratios of the extent of the variability over a period of several minutes, to the average field intensity were observed to be:

at 22 miles with coil antenna 0.25
with vertical antenna zero
at 32 miles with coil antenna 0.44
with vertical antenna 0.07

Ratios of the variability to the maximum field intensity were observed to be:

at 22 miles with coil antenna 0.43
with vertical antenna 0.05
at 32 miles with coil antenna 0.56
with vertical antenna 0.1

Rotating the coil antenna around its vertical axis so as to receive minimum signal showed a variation in the direction of the arriving field as large as 30 deg. over a ten-minute interval.

Observing Location	Miles from Bellefonte	Date	Time	Receiving Antenna	Maximum Fluctuation in per cent	Average Fluctuation in per cent	Direction Shift in degrees
Newville, Pa.	53	Oct. 28	10:15 P.M.	Vertical	38	13	—
" "	53	"	10:35 P.M.	Coil	140	116	62
Andersonburg, Pa.	40.5	"	1:59 A.M.	Vertical	3	3	—
" "	40.5	"	2:19 A.M.	Coil	35	32	30
Lewistown, Pa.	22.7	Oct. 29	6:22 P.M.	Vertical	2	0.5	no observation
" "	22.7	"	6:02 P.M.	Coil	3	0.75	
Sunbury, Pa.	48.5	Oct. 30	4:32 A.M.	Vertical	0	0	no observation
" "	48.5	"	6:22 A.M.	Coil	93	68	
Woodward, Pa.	18.7	Oct. 30	11:59 P.M.	Coil	22	16	8
Hartleton, Pa.	27	Oct. 31	2:12 A.M.	Vertical	18.5	9	—
" "	27	"	1:52 A.M.	Coil	65	57	56
Sunbury, Pa.	48.5	"	6:20 A.M.	Vertical	0	0	—
" "	48.5	"	4:15 A.M.	Coil	21	13.8	15
Hartleton, Pa.	27	"	11:54 P.M.	Vertical	0	0	—
" "	27	"	11:39 P.M.	Coil	12.5	7	12
Woodward, Pa.	18.7	Nov. 1	12:52 A.M.	Vertical	0	0	—
" "	18.7	"	1:07 A.M.	Coil	7	3	2
Sunbury, Pa.	48.5	Oct. 31	9:13 P.M.	Vertical	3	9	—
" "	48.5	"	8:55 P.M.	Coil	38	38	12
Lewistown, Pa.	22.7	Nov. 1	6:10 P.M.	Vertical	8	1.4	—
" "	22.7	"	6:10 P.M.	Coil	30.5	18	—

The coil and vertical antennas were used so that a crude idea might be secured as to the relative extent of apparent shifts in the direction of propagation of the wave, and fading or variation of the received signal strength. The observed shifts of the equisignal zone of the crossed coil beacon signals could be due to either or both effects. The results seem to indicate that fading plays a minor part in the phenomenon.

Observations using an automobile were made again, about November 1, 1927. The results are tabulated as follows, signals from one Bellefonte antenna loop being observed:

The values tabulated in the last column are those maximum shifts observed by following the minimum received signal on a coil antenna through rotation of the coil over a period of several minutes. These data were all taken in mountainous sections of Pennsylvania. They agree in general with the results secured by the first automobile party.

It is hoped that this brief outline of these preliminary observations made by the staff of the Bureau of Standards will serve to focus attention on the performance of this type of radio beacon so that its limitations as a guide for flights at night may be studied and practical information secured whereby these performance characteristics and the conditions influencing them may become better understood.

The data on field intensities given in this paper were secured through the courtesy of Mr. T. Parkinson, Research Associate at the Bureau of Standards, who undertook the automobile trips and made the ground observations.

References to Previous Articles on Radio Beacons for Aircraft

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Applications of Radio in Air Navigation.—J. H. Dellinger. *Engineers and Engineering*, **43**, p. 301; Nov., 1926; *Mechanical Engineering*, **49**, p. 29; Jan., 1927.

Sur les radiophares tournants.—A. Blondel. *Comptes Rendus*, **184**, p. 721; March, 1927.

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OSCILLOGRAPHIC OBSERVATIONS ON THE DIRECTION OF PROPAGATION AND FADING OF SHORT WAVES*

By

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Summary—The short-wave transmission path is generally but not always located in the vertical plane through the transmission and receiving points.

Direction finding depends upon determining the direction of the wave at the receiving point; it does not give accurate results when the twilight zone is in the way of the wave path.

The angle between the earth and the direction of short-wave propagation varies continuously and the changes in this angle are much larger than the changes in angle of propagation in the horizontal plane.

The observations are consistent with the view that the fading is mainly caused by wave interference.

Introduction

THIS paper is intended to be a description of a method for determining the absolute direction of propagation of short waves. The experimental results presented were obtained on 16-meter transatlantic signals. The experiments also give some valuable information on fading.

The method depends upon the beating effects of two received signals, one from the distant and one from a local source when applied to two spaced receivers, the local signal being common to the two. The local signal does not vary or suffer from fading; therefore the beat notes produced in the two sets will represent the signal from the distant station both in relative phase and amplitude, providing the amplitude does not change at a greater rate than can be represented by the beat frequency and this condition is not ordinarily observed. The fading period of the signal[1] was in general five seconds and the beat frequency was held at 500 cycles per second.

The beat note outputs of two receivers are connected to the deflection electrodes of a fiament type of "Braun" or cathode-ray tube and the resulting figure shows the phase difference and the amplitude of the signal waves at the two receiving points. The figure may be a straight line, an ellipse or a circle and it

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[1] Signals from GBK (British beam station) 16m. were generally used.

will change according to the instantaneous value of the amplitude of the field of the signal wave at the two antenna locations.

For a small separation (a fraction of a wavelength) the figure maintained a constant shape and varied only in size. For several wavelengths separation the figure varied continuously in both size and shape indicating random fading and phase relations. Using different antennas, horizontal on one set and vertical on the other, it was found that the phase and amplitude relations between the vertical and horizontal fields generally

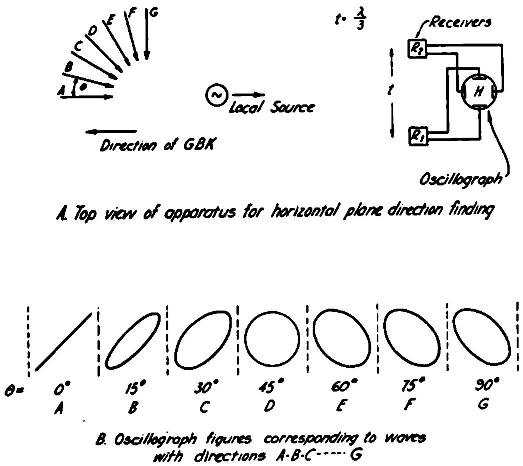


Fig. 1

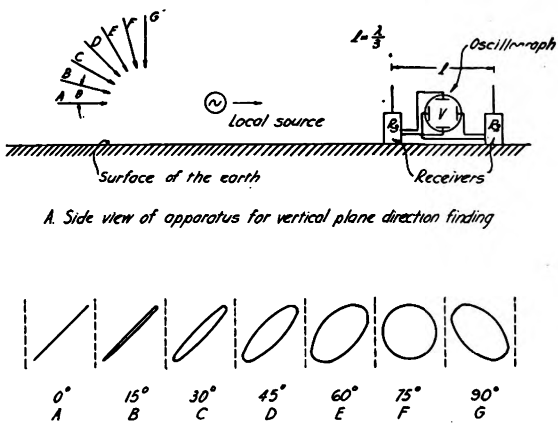
varied at random even when the receivers were close together. In the following description of the experiments the sets were always separated by one-third of a wavelength and only vertical antennas were used. Other distances of separations, for instance one-half wavelength, would have produced similar results.

Direction-Finding in the Horizontal Plane

A top view of the experimental apparatus is shown schematically in Fig. 1-A. The two receivers[2] R_1 and R_2 are located so that the line connecting them is at right angles to the great

[2] Double detection receivers as shown in Fig. 9 of a paper by H. T. Friis and E. Bruce, "A Radio Field-Strength Measuring System," Proc. I.R.E., 14, pp. 507–519, Aug. 1926. The rod antennas were only two feet long in order to prevent distortion of field. The low frequency parts of the sets were made exactly alike so that the oscillograph figure did not change when the beat note was varied. Type 224-A oscillograph made by the Western Electric Co. was used.

circle direction of the transmitting station GBK and the local source oscillator is located in the direction of the transmitting station. Fig. 1-B shows calculated oscillograph figures for different directions, A, B, etc. of propagation of a signal wave. For the direction A there will be no phase difference of the fields at R_1 and R_2 and the oscillograph figure will be a straight line. Direction B will cause a phase difference $(2\pi l \sin\theta)/\lambda = (2\pi \sin\theta)/3$, where l is the separation of the sets and the ellipse B, Fig. 1-B, will be the resulting figure. The figures have been experimentally



A. Side view of apparatus for vertical plane direction finding

B. Oscillograph figures corresponding to waves with directions A, B, C----G.

Fig. 2

checked by means of a second local source oscillator located in directions A, B, etc. in place of a distant signal. The transatlantic signal waves from GBK produce the figures A to C as shown in Fig. 1-B. The size of the observed figures change continuously and they often decrease to zero but there are no appreciable variations in their shape. A straight line or thin ellipse changing in length is the characteristic figure for daylight conditions over the entire transmission path of the signal waves, indicating that the wave path is very nearly located in the *vertical plane of the transmitter and receiver points.* With the local source oscillator located in a fixed position in relation to the two receivers, the whole system has been used successfully as a direction-finder by rotating it until a straight line figure is obtained.

The passing of the "shadow wall" or twilight zone through the wave path of the signal causes large changes in the figure. At times its shape will change continuously from a line to an ellipse, such as C in Fig. 1-B; again it will remain an ellipse for a long period, indicating deviation in the direction of propagation in the horizontal plane. A deviation of as much as 30 deg. from the true direction has been observed. The figure will usually change in amplitude but it is only the largest figures that should be considered in this horizontal direction-finding system. It has unfortunately not been possible to reproduce the actual figures, like a moving picture, but later calculated figures will be shown which may give an idea of the changing oscillograph figures.

Direction-Finding in the Vertical Plane

Fig. 2-A shows the apparatus used. Two receivers R_2 and R_3 and the local source oscillator are now located in the direction of the transmitter. Fig. 2-B shows calculated oscillograph figures for waves A, B, etc. arriving at different angles θ with the earth. A wave A coming along the earth will cause beat notes at R_3 and R_2 with no phase difference, i.e., the figure will be a straight line since both signal A and the local signal travel along the same path. Direction B will cause a phase angle $2\pi l(1-\cos\theta)/\lambda$ and the ellipse B, Fig. 2-B, is the resulting figure. These figures were again checked by means of a second local source oscillator in place of the distant signal. However, this local oscillator was moved around in the horizontal plane and not in the vertical plane as this is much simpler and there should be no difference between a wave propagated horizontally at an angle θ with the direction R_2-R_3 and a wave propagated in the vertical plane of R_2-R_3 and with the angle θ with the earth. For down-coming waves the field at R_2 and R_3 will be the resultant of two waves, a direct wave and one reflected from the ground. The ground being the same at R_2 and R_3 and the sets being at the same height over the ground, the ground reflected waves have no effect on the relative phase of the low-frequency beat notes.

Fig. 2-B shows that small angles cannot be determined as there is too little difference between the ellipse and the straight line. This might be overcome by arranging the sets R_2 and R_3 over each other but complicated effects caused by the reflected waves from the ground are then introduced.

Signals from GBK gave figures *A* to *E* shown on Fig. 2-B. During more than a month's observations the figure would be a straight line during the morning hours, indicating small angle propagation. Towards noon the figure changed continuously from a line to an ellipse, being elliptical most of the time. In the afternoon it changed more rapidly from lines to ellipses. Figures corresponding to angles as large as 60 deg. have been

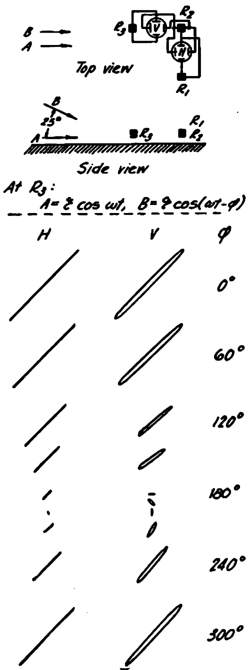


Fig. 3—Wave-Interference Figures.

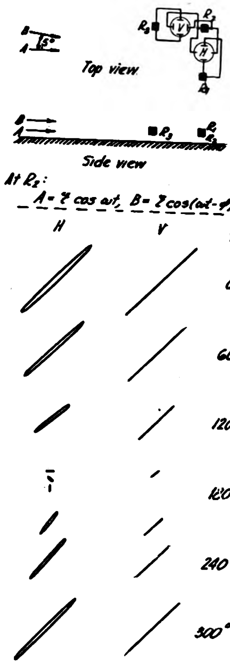


Fig. 4—Wave-Interference Figures.

observed. Later it was found that large angle propagation occurred in the morning also, i.e., the phenomena are very irregular and many further observations are required before definite conditions as to systematic variations can be drawn. So far the results indicate that the changes in angle of propagation in the vertical plane are much larger than the changes in angle of propagation in the horizontal plane. The figures change in size just as in the horizontal direction experiments and here also it is only the largest figures that are to be considered.

The ellipse *B*, Fig. 2-B, tells us that the wave is propagated at an angle θ with the line passing through $R_2 - R_3$, but it does

not identify any particular line in the conical surface satisfying this condition. It is therefore necessary to have both a horizontal plane system and a vertical plane system going at the same time so that figures for both can be observed simultaneously, in which case it is easy to determine the movement of the wave. This was done, one receiving set R_2 being common to the two systems and the two oscillographs being mounted close together.

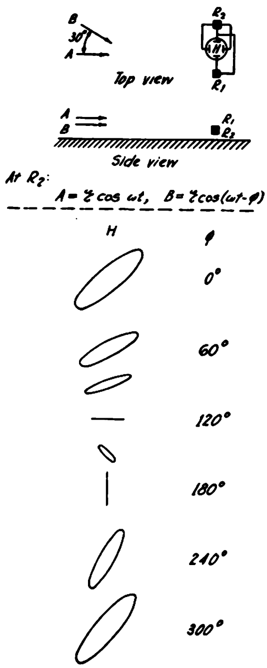


Fig. 5—Wave-Interference Figures.

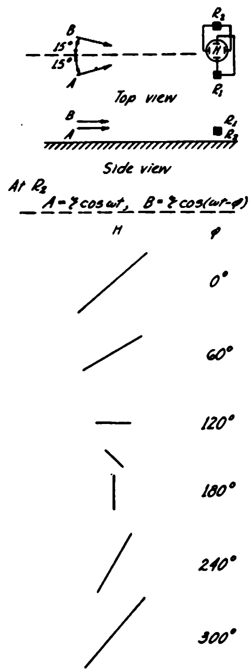


Fig. 6—Wave-Interference Figures.

The results already mentioned are based on the use of this double system.

Fading

The oscillograph pictures mentioned above would continuously increase and decrease in magnitude with an average fading period of five seconds. The small pictures in the fading valleys seemed, at first, to be very irregular, but careful observation disclosed that they practically always consisted of a small line or ellipse rotating quickly one way or the other but practically never exceeding a rotation of more than 180 deg. These char-

acteristic rotating figures, and also the fact that the direction of propagation changes all the time, suggests that fading is caused by wave interference. In Figs. 3, 4, 5, 6, 7 are shown calculated oscillograph figures which would result from a signal composed of two waves A and B of the same amplitude[3] propagated in different directions as shown in the top of the figures. It is assumed that the relative phase ϕ between the waves changes

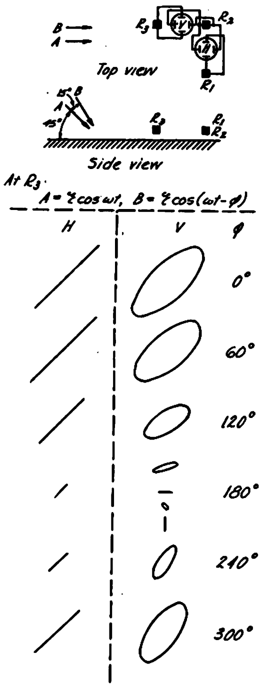


Fig. 7—Wave-Interference Figures.

continuously and the figures correspond to different phase angles. Now these figures have a striking resemblance to the actual figures observed on signals from GBK. The signal figures are sometimes much more complicated but this may be caused by interference between more than two waves. Also the amplitudes of the two waves may not be alike in which case the signal will not decrease to zero during a fading period, as is often found to be the case. The figures shown in Figs. 3 and 4 illustrate the rotating

[3] It is assumed that each of the two waves induce the same e.m.f. in the receiving antenna.

feature of the actual figures. The ellipse figure in Fig. 3 and the line figure in Fig. 4 rotate clockwise with increasing phase angle ϕ. It would have been counter-clockwise in case the angle ϕ had decreased. Note that the speed of rotation is fastest when the figure is small which is in agreement with actual observations. As many as four consecutive fading periods have had the same rotation direction, which means that the relative length of the two wave paths has changed four wavelengths.

In Figs. 3 and 4 the two waves A and B are located in the horizontal plane and the plotted figures correspond to a horizontal plane direction-finding system. In Figs. 5, 6, 7 are plotted the figures that would be observed simultaneously in a horizontal plane direction-finding system, H, and in a vertical plane direction-finding system V. These figures are typical of actual observations.

The total length of each wave path is of the order of 300,000 wavelengths, so it is not remarkable if one path changes a few wavelengths in relation to the other, and thereby causes a continuously changing phase shift of the two waves.

In addition to wave-interference fading there may also be absorption fading, by which is meant quick changes in absorption of a single wave. It is also possible that a single ray may fade due to path changes analogous to what would be produced in light reflection by a rippling reflecting surface. It is believed however that wave interference is by far the most common cause of short-wave fading.

AN AUTOMATIC RECORDER FOR MEASURING THE STRENGTH OF RADIO SIGNALS AND ATMOSPHERIC DISTURBANCES*

By

E. B. Judson

Laboratory for Special Radio Transmission Research, Bureau of Standards, Washington, D.C.)

Summary—*A description is given of apparatus for automatically recording the field strength of low-frequency stations and atmospheric disturbances. The receiver, amplifiers, rectifier, and recorder are switched on by relays controlled by a clock and arranged so that for different 5-minute periods during the hour the strength of several stations may be recorded. The sensitivity of the system remains constant over long periods, provided the filament and plate currents do not change. Calibration can be made at any time from either a radio-frequency or audio-frequency source. Typical curves of the variations in signals and atmospheric disturbances are shown.*

THE recording system described was designed for the purpose of obtaining knowledge of the behavior of radio signals over considerable periods without the presence of an observer. With the apparatus used it is possible to obtain hourly observations of the field strength of several stations and atmospheric disturbances throughout the twenty-four hours.

The entire apparatus is controlled by a clock, which, at different five-minute periods during the hour, actuates a series of relays automatically turning on the receiving set and amplifiers and tuning to the desired stations.

Fig. 1 shows the schematic diagram of the circuit and Fig. 2 shows the arrangement of the apparatus.

The receiving set is of the conventional autodyne type with two tuned circuits having a tuning range from 60 kc. (5000 m.) to 12.0 kc. (25,000 m.). Two antennas are used, one having an effective height of 16 meters for reception of transatlantic stations and atmospheric disturbances, while the other having an effective height of 1.5 meters is used for nearby American stations.

In order to insure greater constancy of sensitivity, only audio-frequency amplification is used. This consists of two

* Original Manuscript Received by the Institute, January 5, 1928. Publication Approved by the Director of the Bureau of Standards of the U. S. Department of Commerce. Read at the meeting of the International Union of Scientific Radiotelegraphy at Washington, Oct. 13, 1927.

stages of transformer coupling followed by four stages of resistance-capacity coupling. With this arrangement, as long as all the filament currents remain constant and the plate voltages do not change the system retains its calibration with sufficient accuracy (within 10 per cent) over periods of several months.

The last audio-frequency amplifier is coupled to a rectifier circuit, containing the recording galvanometer, through an air-core audio-frequency transformer. A three-electrode tube having the plate and grid connected is used for rectification.

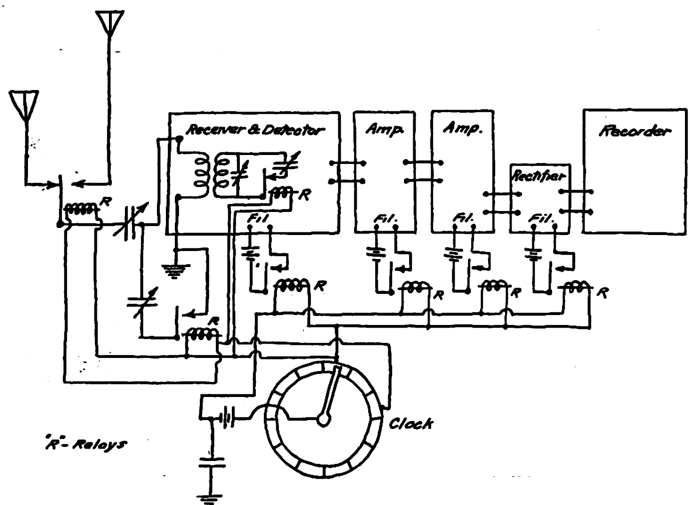


Fig. 1—Automatic Recorder. (Laboratory for Special Radio Transmission Research—Bureau of Standards.)

Generally, tubes used in this manner give a slight galvanometer deflection, when no signal is present, due to the initial velocity of the electrons. This current, however, can be balanced out by a reversed e.m.f. across the galvanometer if it is comparable to the current received from signals.

Records of the variation of signals and atmospherics are made with a Cambridge-Paul Thread Recorder.

This instrument is essentially a sensitive recording galvanometer having a moving coil arranged to give a series of instantaneous records of the galvanometer deflections. The moving coil of the galvanometer has attached to it a pointer which overhangs the drum carrying the paper, while between the pointer and the drum an inked thread is stretched parallel

to the axis of the drum at a short distance above its surface. A presser bar is situated above the galvanometer pointer which is normally held free of the pointer by a cam and its follower. At regular intervals the cam makes a half revolution, first allowing the presser bar to fall upon the pointer and immediately raising it to its normal position. As the presser bar falls it depresses the pointer on the drum nipping the inked thread between the pointer and the paper producing a dot on the paper, which makes a visible record of the deflection of the galvanometer at the moment. The drum of the recorder is arranged for two

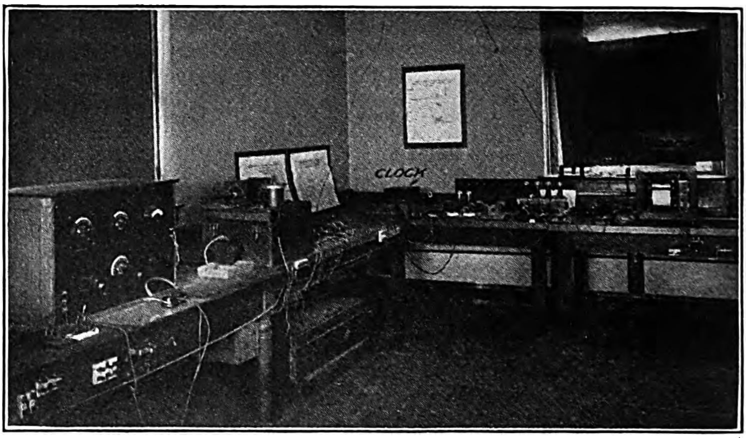


Fig. 2.

speeds, one giving a complete revolution in twenty-four hours and one giving a revolution every two hours for short time records. For the present purpose the twenty-four hour revolution is used.

The clock controlling the entire system has its face plate equipped with twelve brass segments arranged in a circle, each segment representing five minutes. A spring on the end of the minute-hand makes an electrical connection with each segment as it passes over.

The cam allows the presser bar to depress once every 30 seconds, making ten dots on the paper for one five-minute segment.

For the present purpose, the first five-minute segment in the hour is connected so that when the minute-hand contact passes

over it, a relay closes, turning on the receiving set, tuned to the station having the highest frequency of those to be measured. The amplifiers, the rectifier tube, and the presser bar control are switched on simultaneously by means of other relays.

The following segments operate the same relays as the first, but are arranged to close other relays which connect parallel variable capacities across the primary and secondary tuning ca-

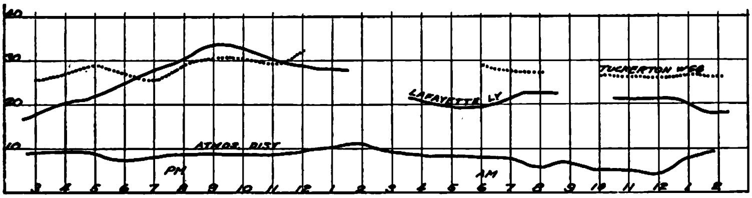


Fig. 3—LY, WGG and Atmospheric Disturbances, March 11–12, 1927.

pacities of the receiving set. This allows the receiver to be tuned to other frequencies effective only for a certain clock segment.

The entire system is slightly sensitive to change of tone. The pitch of the signals being recorded is therefore set to 1000 cycles by comparison with an electrically-driven 1000-cycle tuning fork. A 30-ohm damping resistance across the galvanometer coil slows its period so that it is little affected by change of speed in transmission. Absolute calibration of the receiving

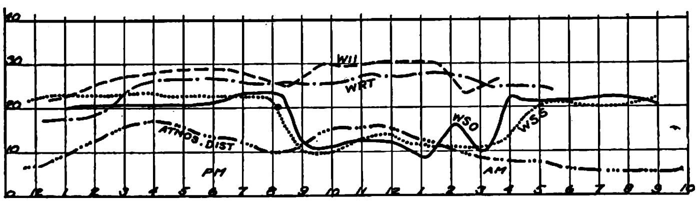


Fig. 4—WSS, WRT, WII, WSO and Atmospheric Disturbances, July 12–13, 1927.

set is obtained from a radio-frequency oscillator feeding into the antenna circuit, while the amplifiers and rectifier may be checked at any time by connecting them to the output of the telephone comparator.[1]

The deflection-current curve of the galvanometer is nearly straight over the usual range of observations. As a protection

[1] Proc. I.R.E., 12, p. 521, 1924.

against heavy atmospherics the deflection is limited just below full scale by limiting the output of the last amplifier tube.

Continuous records of several stations and atmospheric disturbances have been made at the Laboratory for Special Radio Transmission Research of the Bureau of Standards since February, 1927. During the winter months when atmospheric disturbances were low, records were made of Lafayette (LY) and Rugby (GBR) along with Tuckerton (WGG) or Rocky Point (WSS). During the heavy static season, measurements were confined to four American stations, namely, two New Brunswick stations, (WII and WRT), Rocky Point (WSS) and Marion (WSO), reception being on the small antenna. The large antenna, however, is used for reception of atmospheric disturbances.

Fig. 3 shows a typical 24-hour record of Lafayette (LY), Tuckerton (WGG) and atmospherics at 12 kc. (25,000 m.).

In Fig. 4 are given typical 24-hour summer curves, of WSS, WRT, WII, WSO and atmospherics, which show in the case of WSS and WSO an apparent interference of the reflected and direct waves during the night.

DISCUSSION ON THE DISTORTIONLESS RECEPTION OF A MODULATED WAVE AND ITS RELATION TO SELECTIVITY* (F. K. Vreeland)

Lester L. Jones†: Dr. Vreeland's paper has been very interesting to us because it indicates the answer to selectivity problems in broadcast reception. The question of what can be done with the circuits of our old teacher, John Stone Stone, tends to be confused nowadays.

I am one of those who believe that a great deal may be accomplished by proper use of the selectivity inherent in the old

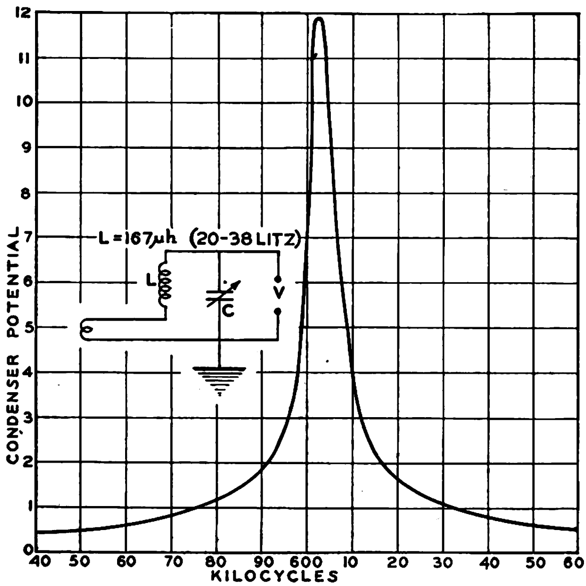


Fig. 1—Selection by a Tuned Circuit.

art circuits. Dr. Vreeland brings up very interesting viewpoints as to the selectivity of tuned circuits. There is one point, however, which appears to me to be somewhat misleading.

In the fifth and seventh paragraphs of the section "Limitations of Selectivity by Resonance," Dr. Vreeland states, "Because of the geometric property of such circuits the ordinates of each graph for a two-circuit system are equal to the squares of the corresponding ordinates for a one-circuit system" and

* Presented at the Annual Institute Convention, January 9, 1928. Published in the PROCEEDINGS, 16, 255, March 1928.

† Consulting Engineer, New York City.

referring to Figs. 1 and 3, that "These graphs are perfectly general and independent of any particular values of inductance, capacitance, or frequency, and they do not involve any assumptions as to whether the circuits are coupled by amplifying tubes or otherwise, or as to the degree of amplification." This statement raises a point which is at variance both with my own experience with coupled circuits and with the impression to be

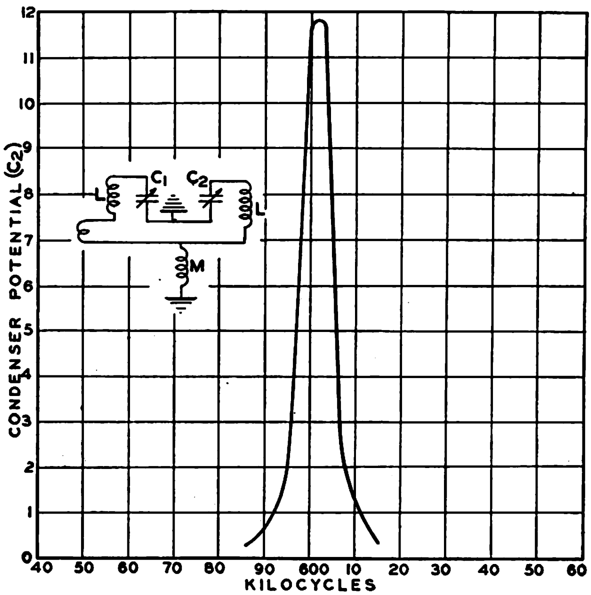


Fig. 2—Case of Optimum Inductance Coupling.

gotten from recent considerations on the question of selectivity. I think the truth is somewhere between Dr. Vreeland's statement and these recent considerations.

In fact, to check up on the theory I had learned years ago with Dr. Goldsmith, I had curves made today using coupled circuits with tube voltmeters for measuring amplitudes and confirmed my belief that coupled circuits of the old art are not nearly so bad as some would have us believe.

If you take two sharply tuned circuits having resonance curves, such as in Fig. 1, and couple them together moderately the curve of response for the second tuned circuit is as shown in Fig. 2. This is taken with the first circuit driven by an oscillator adjusted to the different frequencies shown as the ab-

scissas. You will note that the curve of Fig. 2 is broader than that of Fig. 1 at the top and narrower at the base. It is only when the coupled circuits are coupled so loosely as practically to lose the desired signal that the response curve at the peak is of the form stated by Dr. Vreeland. But why should one couple tuned circuits so loosely as to lose the desired signal? At the base part the response curve for two moderately coupled circuits is hardly different from the geometrically selective response curve for two very loosely coupled circuits. Here, I believe, Dr. Vreeland's statement is correct.

I do not know what constants Dr. Vreeland has for his bridging reactance X_3. It appears to me, though, that his band

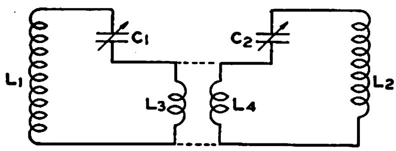


Fig. 3.

selector could easily be redrawn as a pair of coupled circuits as shown in Fig. 3, in which L_1 and L_2 are loading coils, C_1 and C_2 tuning condensers and L_3 and L_4 coupling inductances. By substituting a conductive coupling for the inductive coupling (see dotted line) these coupled circuits have the appearance of Dr. Vreeland's band selector.

In fact, by slighly over-coupling the tuned circuits a response curve having incipient humps and quite square topped could be obtained.

I am sure that if the attention of radio engineers is directed to these desirable characteristics in the old art some very good receiving circuits can be designed.

USE OF AN OSCILLOGRAPH FOR RECORDING VACUUM-TUBE CHARACTERISTICS*

By

W. A. Schneider

(Department of Physics, Washington Square College, New York University)

Summary—*This paper describes the method involved and results obtained in using an oscillograph for plotting photographically vacuum-tube characteristics. The main requirement is an alternating voltage of fairly pure wave-form with as few harmonics as possible. The "dynatron" action of vacuum tubes is shown very clearly when large alternating e.m.f.'s are applied to the grid. Static and dynamic characteristics can also be quite easily recorded.*

THE most common method used in the study of vacuum-tube characteristics is to employ voltmeters and ammeters, and, by connecting them into a suitable circuit, take readings of both instruments, as one of the factors (e.g., E_p) is deliberately varied. After this the readings are recorded in a table, and then transferred to a graph. In this way a picture is obtained of what actually takes place in the tube if we did not have to stop to take readings. The method is, however, long and tedious, especially so if all the various types of characteristics for the many changes which may occur in the study of the operation of a vacuum tube have to be plotted.

The writer was interested in the study of the behavior of vacuum tubes when large positive potentials were applied to the grid of the tube, i.e., when the tube acts as a negative resistance. The characteristics in this case have been studied and interpreted by Hull.[1] The use of the negative resistance action of the vacuum tube has also enabled Barkhausen and Kurz[2] to obtain oscillations experimentally by means of a vacuum tube of enormously high frequency (λ about 10 cm.). Many papers have been published by Gill and Morrell[3] to explain the production of these very short waves.

The oscillograph was used very effectively in plotting these curves. Since the advent of a one-element oscillograph of low price on the market, this instrument has become more and more a piece of general laboratory equipment. Among its many uses

* Original Manuscript Received by the Institute, February 24, 1928.
[1] A. W. Hull, *Phys. Rev.* 7, p. 1, 1916, and Proc. I.R.E., 5, p. 5, 1918.
[2] Barkhausen and Kurz, *Physikalische Zeitschrift*, Jan. 1920.
[3] Gill and Morrell, *Phil. Mag.* S 6, 49, No. 290, Feb. 1925.

such as tracing transient phenomena, wave shapes, current-voltage phenomena, etc., we can now add that of tracing vacuum-tube characteristics of almost any description. The only main requirements are a source of alternating current—frequency not important—of as pure a wave-form as possible. The lower the frequency, the more the characteristic will be spread out on the time axis. The writer after trying many different sources found that the wave-form as supplied in New York City is nearly enough sinusoidal so as not to affect the picture noticeably as recorded on the photographic drum. The harmonics may show up when large magnification is applied. If the power supply is not of good wave-form, a contact maker on a rotating slide wire, properly connected, will furnish the necessary variable e.m.f., this arrangement having the advantage that the frequency can be changed at will.

The only other requirement is that the oscillograph element shall have the necessary sensitivity. For most vacuum-tube work a sensitivity of 10^{-4} amperes per mm. deflection on the rotating drum is sufficient, although it is not difficult to obtain elements of much larger sensitivity.

Although the curves shown in the diagram were taken with a 3-element Siemens and Halske Instrument, a cheaper one-element oscillograph will suffice for this purpose just as well.

The curves shown in Fig. 1 were taken to show the "dynatron" action of an ordinary UX-201A vacuum tube. The plate voltage E_p was kept at a fixed value and to the grid was applied the alternating e.m.f. of various values (10 different values were used from 28.2 volts for curve I to 155 volts for curve 10) as shown in Table I.

TABLE I

Curve	Max. E_g (volts)	Curve	Max. E_g (volts)
1	28.2	6	98.7
2	42.3	7	112.8
3	56.4	8	126.9
4	70.5	9	141.0
5	84.6	10	155.0

The negative characteristic is very evident. The different pictures in Fig. 1 are for various fixed values of plate voltage. The graphs in each case must be read from the right to the left on account of the direction of rotation of the film and it will be further noticed that only one half is necessary to give the actual characteristic.

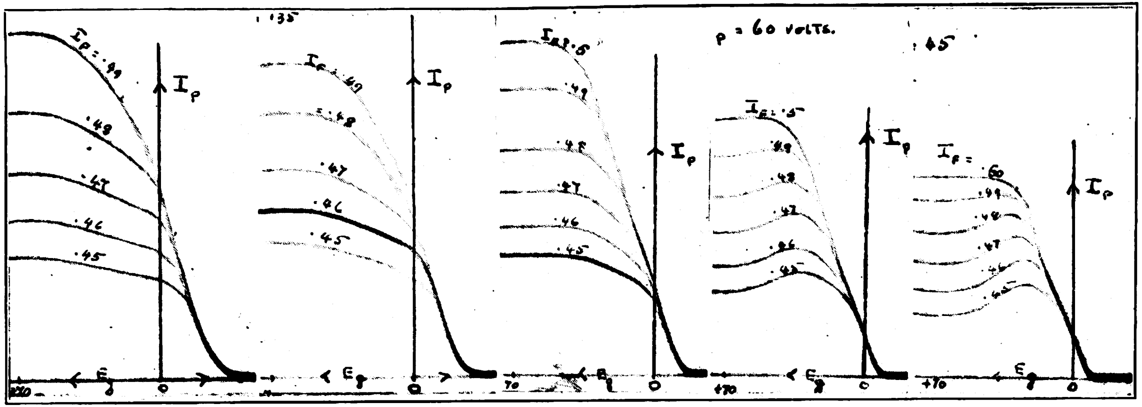


Fig. 3

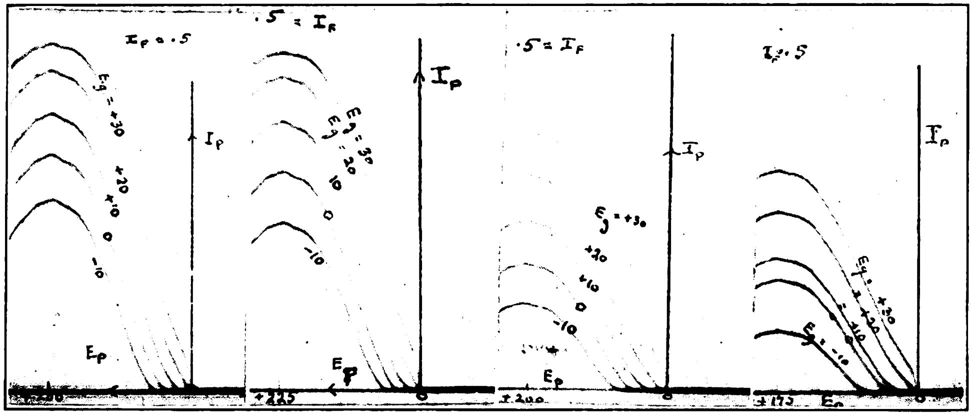


Fig. 4

Although the shape of the curves in Fig. 1 is interesting they can not be interpreted very easily because the units on the grid-voltage (E_g) axis are different for each curve, the maximum value on the figure representing all the values given in Table I.

In Fig. 2 this difficulty is overcome by photographing the various I_p–E_g characteristics under conditions of different plate voltages, the grid-voltage axis now being the same scale for all the curves in the same diagram, the different diagrams being plotted with grid values of 120, 90, 60, 30, and 20 volts respectively.

Very many interesting results can be seen from these curves. One of these is that the current through the tube under certain conditions will be larger from plate to filament than from the usual filament to plate. It is not the purpose of this paper, however, to go into the results found.

The next set of curves represented in Fig. 3 show the effect of changing the filament current in a plate-current, grid-voltage characteristic. The set of curves obtained in each picture of Fig. 3 are for a constant plate voltage. The different sets are for plate voltages of 165, 135, 90, 60, and 45 volts respectively, the variation in grid voltage in each case being between limits of ± 70 volts. The filament current was varied from 0.45 amperes to 0.5 amperes (for a UX-171).

In Fig. 4 will be seen the effect of grid voltage on displacement of the characteristic with respect to the plate-current axis. They are all E_p–I_p curves, each set being for a fixed filament current, but 5 different values of grid bias, viz. $+10$, 0, -10, -20, and -30 volts. Here again most interesting results are noticed. In all curves for a grid bias of -30 volts the characteristic takes a sudden dip at a small positive plate potential in such a fashion as just to give zero plate current when the variable plate voltage is zero. All these effects fall in line exactly with other methods and theory.

The above curves are typical of the results to be obtained by using the oscillograph for this purpose. Many other similar uses in connection with characteristics naturally suggest themselves, such as transformer characteristics, etc. In Fig. 5, for example, are curves showing the relation between static and dynamic characteristics. They are plate-current, grid-voltage characteristics (for constant I_F and E_p). The highest curve is the static characteristic. The other five lower curves are for resistances

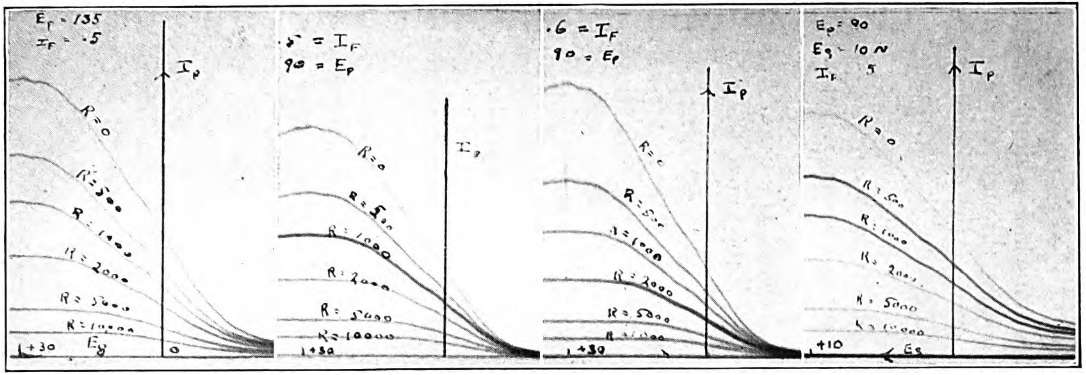

$E_p = 135$
$I_F = .5$
I_p
R=0
R=500
R=1000
R=2000
R=5000
R=10000
+30
E_g
0
.5 = I_F
90 = E_p
I_p
R=0
R=500
R=1000
R=2000
R=5000
R=10000
+30
.6 = I_F
90 = E_p
I_p
R=0
R=500
R=1000
R=2000
R=5000
+30
$E_p = 90$
E_g - 10 N
I_F .5
I_p
R=0
R=500
R=1000
R=2000
R=5000
R=10000
+10
E_g

Fig. 5

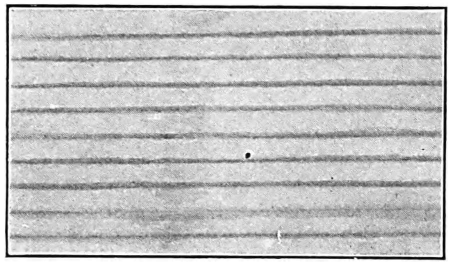


Fig. 6

in the plate circuit of 500, 1000, 2000, 5000, and 10,000 ohms respectively. The slopes of these curves play important roles in the operation of the tube in any type of vacuum-tube circuit.

Lastly in Fig. 6 is shown a calibration which is simply carried out by passing certain known currents through the element (in this case about 2 milliamperes) showing the uniform sensitivity for all parts of the film and at the same time giving a permanent record of the sensitivity of the element.

When it is remembered that each of the above curves are drawn in 1/120th of a second the advantages of such a method are obvious.

In conclusion the writer wishes to thank Mr. Frank Wenger for valuable aid rendered in carrying out the above tests.

BOOK REVIEWS

The Theory of Sound, BY LORD RAYLEIGH. Macmillan and Company, London. Revised Edition, 1926, 494 pages. 15 schillings net.

BY HARVEY FLETCHER†

"The Theory of Sound" by Lord Rayleigh has been the standard text on the subject during the last 50 years. It has been used to good advantage by investigators interested in the transmission of electrical vibrations. It is an interesting coincidence that the first edition of this book appeared in 1877, the same year that Alexander Graham Bell demonstrated his invention of the telephone before the British Association for the Advancement of Science. The previous year Lord Kelvin had seen Bell's demonstration at the Centennial Exposition in Philadelphia and had made reference to it in a lecture that fall at Glasgow. It is remarkable that a text which is still invaluable to a telephone engineer was written before the invention of the telephone. To one not familiar with the book, this may seem incredible.

Lord Rayleigh's treatment of the subject of vibrating systems was so thorough that when the art of electrical transmission of speech was developed, the equations which he had already written down proved to be adequate for the phenomena of this novel field provided only that the meanings of the constants were properly altered. Nowadays the tendency is in the opposite sense for most engineers are more familiar with electrical transmission than acoustical transmission. Consequently, when a problem involving the transmission of mechanical or acoustical vibrations presents itself, the analogous electrical circuit is sought so that one may better understand the process of transmission.

A second edition of "The Theory of Sound" appeared in 1894. It was greatly revised and enlarged over the first edition and contained among other additions a new chapter on "Electrical Vibrations." In this chapter it was shown that the equations which were formerly developed for representing mechanical motions could be applied directly to the new problems of electrical transmission. In recent years an owner of the two volumes of this edition spoke of his ownership with considerable pride as it was very difficult to obtain copies. In 1926, however, it was reprinted and now any one may obtain copies of this valuable work.

† Bell Telephone Laboratories, Inc., New York City.

In the introduction a general description of the phenomena of sound is given. The material in most elementary text-books on Physics is patterned very largely after that presented in its pages. It includes an account of experiments on the speed of sound in liquids and solids and a statement of the factors which govern the physical characteristics of musical sounds; namely, pitch, loudness, and quality. The nature of the various musical scales is explained in the way which has since become customary.

The author then proceeds to give the simple theory of harmonic vibrations and applied it to systems having a single degree of freedom. In his treatment of the so-called intermittent vibrations he laid the foundation for the theory of a modulated carrier current wave which is now so familiar to radio engineers. The equations developed in this connection can be applied directly to the modern problems of modulation. In his treatment of the asymmetrical natural vibrations of one degree of freedom, he derives equations which are applicable to the case of an overloaded vacuum tube or to a rectifier.

Next comes the development of the theory of a vibrating system having any number of degrees of freedom. The sets of linear differential equations used for representing the behavior of such a system are the same as those used for representing the behavior of a connected system of electrical networks; consequently, the theory expounded by Rayleigh can be transferred readily to the analysis of the properties of electrical systems such as wave filters, artificial lines, and other complicated networks. The only alterations required are those to be made in the meanings of the various constants. One is astonished to observe how completely the theory developed in this book parallels the more recent generalized theory of electrical networks.

The reciprocal theorem is proved for a very general case. It is shown that it holds for impulses as well as for steady-state frequencies. As an illustration of this theorem, the case of the struck piano wire is cited. If the hammer strikes the wire at a point X_1 the displacement at a point X_2 is the same as the displacement would be at X_1 if the stretched wire were struck at the point X_2. One familiar with electrical circuits can very readily apply the equations to a corresponding electrical case. There is an interesting application of this reciprocal theorem in the case of telephone booths. If two people, one inside such a booth and one outside of it, talk to one another with voices of equal strength,

they hear one another equally well, provided of course there are no other sounds present.

It is well to give a word of warning here. There are frequent misapplications of this reciprocal theorem because the meaning of sources of equal strength is often misunderstood. For example, in fluids sources of equal strength are those "produced by the periodic introduction and abstraction of equal quantities of fluid or something whose effect is the same." This does not necessarily mean that the same amount of sound power is radiated from two sources of equal strength. "For instance, a source close to the surface of a large obstacle emits twice as much energy per second as an equal source situated in the open."

In his treatment of the transverse vibration of strings both when free and when loaded, Rayleigh laid the foundation for the theory of transmission of electrical vibrations on loaded and non-loaded cables. The general equations for determining the arbitrary constants from the initial conditions of the system, which he developed for the acoustic case, are equally valid for the electrical problem.

The next subject is the vibration of bars. For the longitudinal and the torsional oscillations, Rayleigh shows that the same equations hold as for the transverse oscillations of stretched strings, provided that the constants which in the earlier case stand for the tension and linear density of the string are in this case interpreted as the elasticity and the volume-density of the material of the bar. In the author's characteristic language: "A bar under tension which is sufficient to double its length will emit the same note as that due to longitudinal vibration."

The transmission of vibrations from one bar to another of different material and size across a mechanical junction corresponds quite closely to the transmission of electrical vibrations from one circuit to another with different constants. It is strikingly difficult to transmit energy of vibration from air to steel, or vice versa, for the amount which crosses the junction is only 0.00001 of that which arrives at it. In other words, a transmission loss (sometimes called reflection loss) at a junction between air and steel is about 50 TU. It is shown that the free period of torsional vibrations of bars must lie between 1.4 and 1.7 times the longitudinal free period.

The treatment of the lateral vibration of bars leads to much more complicated differential equations, but the subject is

treated in a masterful fashion. He shows that *the period of vibration of a solid object of any shape whatever varies as the linear dimension if the material and shape are kept constant.* As one of the practical applications, it is shown that the period of vibration of a tuning fork is independent of the thickness perpendicular to the plane of bending, inversely proportional to the thickness in the plane of bending, and proportional to the square of the length.

The treatment of the vibration of stretched membranes leads to a differential equation which is the same as for a stretched

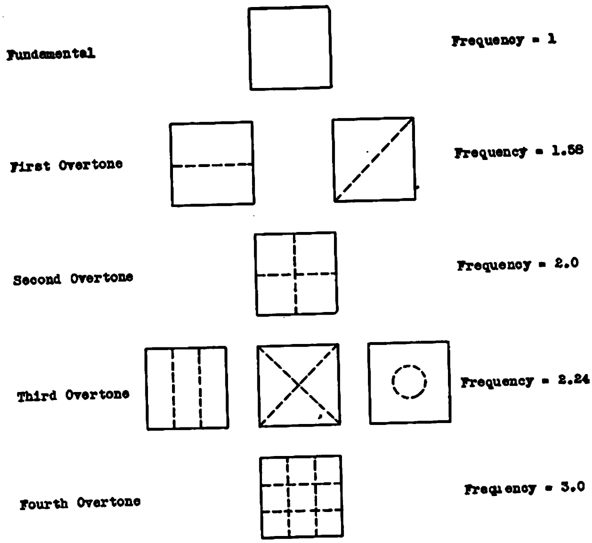


Fig. 1

string except that an additional term is added. When the boundary is rectangular, frequencies are given by the equation

$$f=\sqrt{\frac{T}{\rho}}\sqrt{\frac{m^2}{a^2}+\frac{n^2}{b^2}}$$

where T is the tension, ρ the superficial density, a and b the dimensions of the rectangular boundary, and m and n the integers. When the boundary is rectangular the solution leads to a Fourier series and when the boundary is circular to a series of Bessel functions. It is interesting to note the modes of motion associated with the harmonics of such a stretched membrane. The sketches

in Fig. 1 indicate the forms for the first four overtones of the square membrane.

Similarly, in Fig. 2 are shown the frequencies and modes of vibration for a circular boundary. It is seen that most of the periods are inharmonic so that a sound emitted from such a membrane would not in general be musical. However, the author

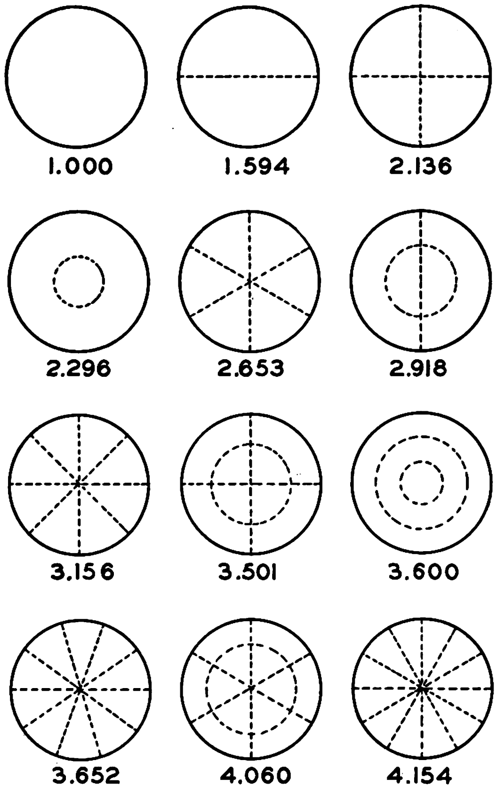


Fig. 2

points out the interesting fact that the first, the second, the fourth, and the sixth overtones form a series of musical intervals which are approximately 4/3, 5/3, and 2, that is, notes on the major musical scale. A circular membrane has a fundamental frequency 4 per cent lower than that of a square membrane of the same material, equal area, and equal tension.

In the treatment of the vibration of plates the mathematical difficulties are still greater and only approximate solutions can be

obtained. However, they prove to be sufficiently accurate to check with what experimental values are available. Rayleigh treats the case of a vibrating ring and from the results is able to make some interesting deductions concerning the pitch of bells.

As I said before, the last chapter of the first volume deals with electrical oscillations; the important formulas used in alternating-current theory are deduced, including coupled circuits and a detailed account of the induction bridge. The author shows how a branch containing a self-inductance and a resistance may be treated mathematically as though it were a single resistance. In this connection it may be interesting to remark that Lord Rayleigh frequently refers the reader to Heaviside's work for more complete solutions of the electrical problems dealt with. It will be remembered that this second edition came out in 1894. His frequent reference to Heaviside's work indicates his high regard for it even at this early period. Inasmuch as this book appeared in its second edition six years before the invention of the loading coil, it is interesting to quote the following paragraph:

> For a further discussion of the various cases that may arise the reader must be referred to the writings of Heaviside already cited. The object is to secure, as far as may be, the propagation of waves without alteration of type. And here it is desirable to distinguish between simple attenuation and distortion. If, as in (16) and (18) P is independent of p, the amplitudes of all components are reduced in the same ratio, and thus a complex wave travels without distortion. The cable formula (15) is an example of the opposite state of things, where waves of high frequency are attenuated out of proportion to waves of low frequency. It appears from Heaviside's calculations that the distortion is lessened by even a moderate inductance.
>
> The effectiveness of the line requires that neither the attenuation nor the distortion exceed certain limits, which, however, it is hard to lay down precisely. A considerable amount of distortion is consistent with the intelligibility of speech, much that is imperfectly rendered being supplied by the imagination of the hearer.

The second volume is concerned principally with a particularly detailed study of aerial vibrations. The general equations representing the motion in a disturbed liquid or gaseous medium are first developed. These equations are then applied to several important cases. The first is the simple case of the vibration of air in a tube. The equation for representing such a disturbance is the same as that obtained for the transverse vibration of strings, or the torsional or longitudinal vibration of solid bars. These three types of vibrators produce sounds with

overtones harmonically related to the fundamental. For this reason they are usually employed in making musical instruments.

Rayleigh shows that the transference of energy along a wavetrain of plane sound waves can be described by equations with constants formally identical with those used in circuit theory. To take an example; there is a quantity, the product of the density of a fluid by the speed of sound in it, which enters the equations in precisely the same manner as electrical resistance into the circuit equations, and is in fact designated as "radiation resistance." The rate of flow of energy through a unit area parallel to the plane of the wave-fronts is the product of this quantity by the square of the particle-speed. Analogous formulas for other forms of Ohm's law also hold—the velocity of the particles corresponding to the electrical current, the amplitude to the quantity of electricity flowing in the circuit, and the pressure change to the electrical potential difference. The formulas for the velocity of propagation of sound in terms of the gas constants are developed along the usual lines; in addition it is shown that an exact solution of the problem requires that different frequencies be propagated at different rates. However, the differences are very small and are usually negligible.

From his equations the author predicts that an object placed in a fluid traversed by sound waves must experience a pressure —in certain conditions, a torque—the amount of which can be computed from the intensity of the sound. This was the origin of the "Rayleigh disk" (the qualitative theory is given in this book, the quantitative theory being first worked out by Koenig) which is now used in the fundamental work of calibrating instruments to measure sound intensities.

In connection with the development of the theory of vibration of air columns, he deduces from the theoretical equations the end correction of the tube, i.e., the amount by which its length departs from one-quarter or one-half of the wavelength of its fundamental according as it is closed at both ends or only at one. The conditions at the boundary between two media having different constants are considered and equations deduced which give the amount of reflection and refraction of a sound beam striking such a boundary at any angle. As a practical application, it is shown that regular corrugations on a reflecting plane do not impair the reflection of sound unless their height is comparable with the wavelength. Considering the size of the

wavelengths of ordinary sounds, one sees that even a wall which seems very rough to eye or hand will reflect sound like a perfect mirror.

Next the author deduces the general equations for spherical waves and makes some important applications, notably to the theory of conical horns and ear trumpets; this problem involves the reciprocal theorem which I have already mentioned. It is shown that a listener will obtain the same loudness when holding a conical horn to his ear as he would obtain if the horn were inverted and its small end held at the speaker's mouth. This conclusion, however, is correct only under certain conditions to which the connection between ear, horn, and mouth must conform. He points out the definite limitations of a horn radiating sounds having wavelengths larger than the opening of the horn. Similarly, Rayleigh shows why it is that no appreciable shadows are cast by solid objects in the path of sounds usually encountered, and also shows why concave mirrors for concentrating sound are of no avail except when they are very large compared to the wavelength. Concerning the acoustics of buildings, this very significant paragraph occurs:

> In connection with the acoustics of public buildings there are many points which still remain obscure. It is important to bear in mind that the loss of sound in a single reflection at a smooth wall is very small, whether the wall be plane or curved. In order to prevent reverberation it may often be necessary to introduce carpets or hangings to absorb the sound. In some cases the presence of an audience is found sufficient to produce the desired effect. In the absence of all deadening material the prolongation of sound may be very considerable, of which perhaps the most striking example is that afforded by the Baptistery at Pisa, where the notes of the common chord sung consecutively may be heard ringing on together for many seconds. According to Henry it is important to prevent the repeated reflection of sound backwards and forwards along the length of a hall intended for public speaking, which may be accomplished by suitably placed oblique surfaces. In this way the number of reflections in a given time is increased, and the undue prolongation of sound is checked.

From the equations for the transmission of sound through air of uneven temperature, it is clear that sound must be refracted upward through air of which the temperature increases with height and consequently that a speaker's voice must be heard more clearly by a listener above than by one below him.

Rayleigh gives a thorough treatment of the behavior of air resonators and deduces formulas which have since been checked by experiment and have proved very useful in the design of

resonators and other acoustic apparatus involving air chambers. Before leaving the subject of air vibrations in tubes, he considers the various methods of exciting the natural frequencies, including the theory of the excitation of organ pipes and whistles, reed instruments such as the clarinet and the oboe, the Rijke tube which is excited by a hot flame, etc.

In considering the reflection of sound, some formulas are developed which have important application in the design of radio and telephone apparatus involving vibrating diaphragms. Formulas for the extra mechanical impedance which the surrounding air imparts to a circular diaphragm supposed for mathematical convenience to be cut out of an infinite plane are given in terms of functions which may be readily calculated for any numerical case. Formulas are likewise given for the corresponding quantity in respect to a solid sphere oscillating in air. When the sphere is small compared to the wavelength, the effective mass added by the air is one-half of the mass of the air displaced by the sphere. The resistance due to the air is very small and consequently very little sound energy is radiated by such a vibrating device. For example, for a sphere one centimeter in radius vibrating at 1000 vibrations per second, the resistance is only approximately 1/1000 of the mass of displaced air while the mass reactance is 500 times this value. The mechanical impedance offered by the air to the vibration of a small circular disk in an open space is greater than for a sphere, the mass reactance being about 25 per cent greater than for a sphere having the same radius.

The equations given for calculating the intensity near a solid sphere due to a distant source may be used to good advantage in calculating the intensities near the human head, an important use when considering the theory of binaural location of sounds. The air reaction on other types of vibrating bodies, including the string, is calculated. For example, it is shown that the air reaction on the vibrating string is so small that the energy directly radiated is less than 1/40,000 of that radiated when it is attached to a sounding board. Similarly, wires or small objects of any shape offer little obstruction to the passage of sound waves.

When the viscosity of the air is taken into account the equations indicate that the decline in amplitude of advancing sound waves is very small. It is fastest for the high frequencies. For

example, a sound having a wavelength of one centimeter loses two thirds of its initial amplitude in travelling through 88 meters, while a sound of 10 centimeters advances through 8800 meters before suffering an equal reduction. In this treatment important formulas are developed which give the velocity and the attenuation of a sound wave travelling in a tube. These equations have since been shown by direct experimental tests to be valid over wide ranges of frequencies and sizes of tubes. For example, a tube 1.4 inches in diameter has an attenuation of 0.15 TU per foot for a sound of 1000 cycles. In other words, such a sound would be reduced to about 3 per cent of its original intensity in travelling a length of 100 feet of tube. The attenuation for other frequencies is proportional to the square root of the frequency.

The author next develops, in the familiar way, the equations for water waves of two distinct types (gravity waves and capillarity waves). He then discusses at great length the behavior of jets of water under various conditions. One of his interesting conclusions is that a jet having a circular cross-section is the only one which will maintain its form. If a jet issues from an orifice with a non-circular cross-section, its form as it proceeds through the air will vary cyclically about the circular form. Jets of air are also considered and it is shown how to construct a flame which is very sensitive to high-pitched sounds. It is interesting to note that such a sensitive flame is affected most at those points of a standing wave where the ear indicates there are silences. In other words, the ear indicates a maximum loudness of sound at those places where the pressure changes are greatest, whereas the sensitive flame becomes disturbed most at those places where the pressure changes are zero and the velocity of the air particles is at a maximum.

The last chapter is concerned with a description of the facts and theories of audition. Although this portion of the work has lost much of its value owing to the quantity of data subsequently assembled, it shows how competently the author judged the significance of such data as were available when he wrote. He also stresses strongly the need of more accurate experimental data.

As would be expected, he gives a clear presentation of the correct theory of hearing as well as the correct theory of the formation of the vowel sounds. This statement probably would

be disputed by some investigators now working in this field, but I think only by those who do not have a clear notion of the dynamics involved. The additional experimental facts have made it necessary to modify somewhat the views which he has expressed. However, the clear-cut fashion in which he expounds both of these theories is quite in contrast to the confused thinking of many who have written on this subject since his time. Papers are still appearing which show that the authors do not grasp the important points so ably discussed by Lord Rayleigh. For example, he expounds the Helmholtz resonance theory of hearing with the understanding that the pitch is determined from the position of the maximum response, realizing full well that a considerable portion of the end organ must be stimulated by every tone. He also gives an explanation of the production of the summation and difference tones which is undoubtedly correct; namely, that they are due to the non-linear response of the hearing mechanism of the middle ear.

After his clear explanation of the relationship between the two rival theories of vowel production, namely, that championed by Willis and sometimes called the inharmonic theory, and that championed by Wheatstone and Helmholtz and sometimes called the harmonic theory, it is hard to see why there has been such a discussion of the relative merits of these two theories since his book was published. He plainly indicates that these two theories are only different ways of looking at the same phenomena and each is convenient and useful according to the purposes of the problems at hand.

It is frequently stated that when a treatise on a scientific subject has become 10 or 15 years old, it is ready for the cellar or the garret, its obsolescence being due to the rapid advances which science is making. This book on "The Theory of Sound" is certainly an exception. It is now more than 50 years old and it will continue to be used for a good many years to come as one of the principal sources of information concerning the theoretical aspects of the production and transmission of sounds and will contribute largely to all fields of work where vibrations are concerned.

Cunningham Tube Data Book. E. T. CUNNINGHAM, INC., New York City. 84 pages, 9 x 12. Price $2.50.

This book presents in compact form much necessary information on current types of radio tubes used in receivers and socket

power devices. There are brief pointed chapters on radio tube operation and performance with some special reference to the thoriated filament. But the principal value of the book is the detailed description of each of eighteen types of Cunningham radio tubes and the curves and data showing average characteristics of each of these tubes. With the multiplicity of tube types on the market at present, and as other manufacturers make similar lines, this book is a valuable aid in the selection and operation of a tube for best performance, and in the design of equipment to use with the tube.

S. S. Kirby

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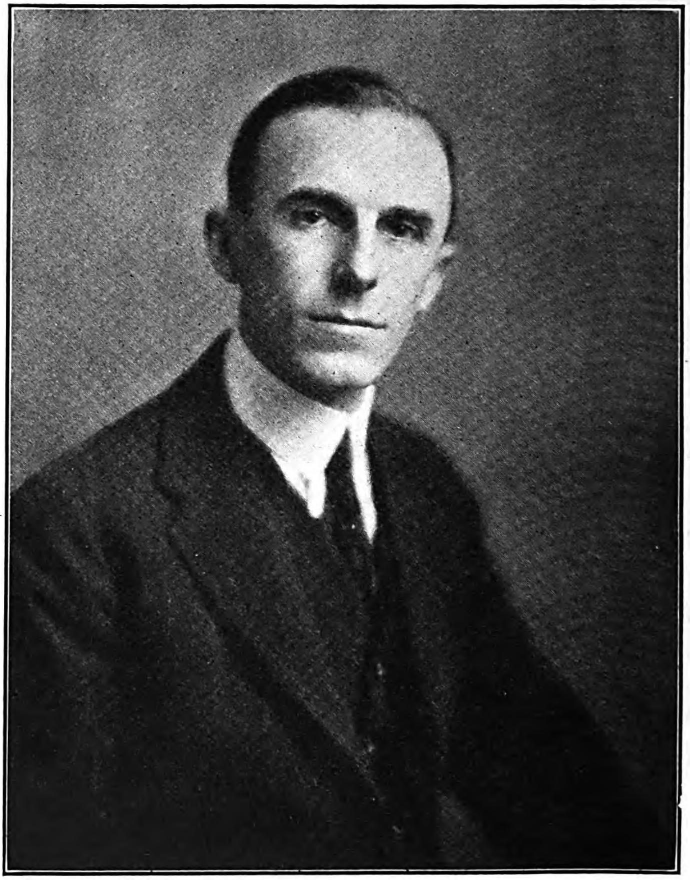

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Member of Board of Direction of the Institute, 1928

Ray H. Manson

MEMBER OF THE BOARD OF DIRECTION OF THE INSTITUTE, 1928

Ray H. Manson was born in Bath, Maine, on August 25, 1877. He graduated from the University of Maine in 1898, and received the E.E. degree from the same institution in 1921. The year following graduation he was an assistant in the Electrical Engineering Department of the University of Maine.

In 1899–1900 Mr. Manson was employed in the Telephone Manufacturing Department and the Electrical Laboratory of the Western Electric Company in Chicago. From 1901 to 1904 he was in the Engineering and Sales Departments of the Kellogg Switchboard and Supply Company at Chicago.

During the period of 1904 to 1916 Mr. Manson was connected with the Dean Electric Company and its successor, the Garford Manufacturing Company, at Elyria, Ohio, the latter four years being Chief Engineer. Since 1916 he has been Chief Engineer of the Stromberg-Carlson Telephone Manufacturing Company at Rochester, New York.

During his association with the manufacturing business over fifty U. S. patents on telephone, phonograph, and radio subjects have been granted him.

Mr. Manson has been active in radio standardization work in the Radio Division of the National Electrical Manufacturers' Association. He is Chairman of the Technical Committee of that organization. For several years he has been Chairman of the Electro-Acoustic Subcommittee on the I. R. E. Committee on Standardization.

Mr. Manson was appointed a Manager of the Institute by the Board of Direction in January of 1927, and was elected by the Membership to the Board for a three-year period on January 3, 1928. He is a Member of the Institute, a member of the American Institute of Electrical Engineers, and a Member of the Society of Automotive Engineers.

CONTRIBUTORS TO THIS ISSUE

Dahl, O.: (See PROCEEDINGS for March, 1928).

Espenschied, Lloyd: Born April 27, 1889 at St. Louis. E.E. course at Pratt Institute, 1909; amateur experimenter and radio operator on ships, 1905–09; assistant engineer Telefunken Wireless Telegraph Company of America, 1909–10; engineer, Department of Development and Research, American Telephone and Telegraph Company since 1910. In charge of transmission of high frequency in wire carrier and radio systems. Responsible for a number of inventions relative to communication systems, both wire and radio; also in field of railway signal and speed control systems and phonograph systems. He has contributed frequently to the PROCEEDINGS; is a charter member and Fellow of the Institute, and served on the Board of Direction of the Institute from 1913 to 1927.

Nelson, J. R.: Born on October 27, 1899 at Murray, Utah. Power Inspector, Western Electric Company, 1922–23; B.S. in E.E., University Southern California, 1925; Engineering Record Office, Bureau of Power and Light, Los Angeles, Calif., 1925; Radio Test, Radio Development Laboratory and Tube Research Laboratory, 1925–27. M.S. in E.E. degree, Union College, 1927. July, 1927 to date, Engineering Department, E. T. Cunningham, Incorporated. Associate member of the Institute.

Owens, Raymond B.: A.B. degree Randolph-Macon College, 1917. Post-graduate work at George Washington University. In naval radio service 1918–19; Radio Test Shop, Washington Navy Yard, 1919–21; Naval Radio Laboratory, Bureau of Standards, 1921–23; Research work in connection with precision apparatus for Naval frequency standardization, Naval Research Laboratory, 1923 to date. Member of the Institute.

Pickard, Greenleaf W.: (See PROCEEDINGS for May, 1928).

Prince, D. C.: Received B.S. in E.E. degree at University of Illinois, 1912; M.S. in E.E., University of Illinois 1913; General Electric Company, Test and Railway Motor Department, 1913–14; State Public Utilities Commission of Illinois, Assistant to Electrical Engineer, 1914–17; Ordnance Department, A.E.F., 1917–19; Research Laboratory, General Electric Company, 1923 to date. Fellow, American Institute of Electrical Engineers and American Physical Society.

Terman, Frederick Emmons: (See PROCEEDINGS for April, 1928).

Turner: H. M.: Born at Hillsboro, Illinois, July 20, 1882. Received B. S. degree, University of Illinois, 1910; M. S. degree, 1915; assistant instructor in electrical engineering, 1910–12. Instructor, University of Minnesota, 1912–18. Assistant professor, Yale University, 1918–26; Associate professor, 1926– . Frequent contributor to the leading engineering journals. Member I.R.E., A.I.E.E., and International Union of Scientific Radiotelegraphy.

Tuve, M. A.: Born on June 27, 1901 at Canton, South Dakota. Amateur radio operator, 1915–21; Received B.S. in E.E. degree and M.A. degree from University of Minnesota; Ph. D. degree from Johns Hopkins, 1926; former Instructor in Physics, Princeton University and Johns Hopkins University. At present, associate physicist, Department of Terrestrial Magnetism, Carnegie Institution of Washington.

Van Dyke, K. S.: Born at Brooklyn, N. Y., December 8, 1892; Received B.S. degree from Wesleyan University, 1916; M.S. degree, 1917. Ph.D. degree, University of Chicago, 1921. Engineering Department, American Telephone and Telegraph Company, 1917–19; Assistant in physics department, University of Chicago, 1919–21; Assistant Professor of Physics, Wesleyan University, 1921–25; 1925 to date Associate Professor of Physics, Wesleyan University. Member of the Institute. Secretary (1927) and Vice-Chairman (1928) Connecticut Valley Section of the Institute.

Williams, N. H.: Ph.D. degree, 1912. Research Laboratory, General Electric Company, 1923–24, engaged in development of screen-grid tube and measurement of the charge of the electron by the "shot-effect." On staff of Physics Department, University of Michigan, 1909 to date. At present Professor of Physics.

Worrall, Robert H.: Educated at Lowell Textile School and George Washington Unversity. With Manchester Traction and Power company, Manchester, New Hampshire, 1912–16 in charge of electrical machine shop; assistant in charge of Transmitter Test Laboratory, Marconi Wireless Telegraph Company, 1916–17; in charge of Receiver Test Laboratory, Washington Navy Yard, 1917–23; research and design pertaining to frequency standardization, Naval Research Laboratory, 1923 to date. Member of the A. I. E. E. and Associate member of the Institute.

Yagi, Hidetsugu: Born, January 28, 1886 at Osaka City, Japan. Graduated and received degrees from Tokio Imperial University, 1909, and Department of Education, Japanese Government, 1919. Lecturer at Sendai Higher Technical School, 1909-10; Professor in electrical engineering, Sendai Higher Technical College, 1910-1912; travelling Fellow, Department of Education of Japanese Government, studying in Berlin, Dresden, London, and at Harvard, 1913-16; Professor of electrical engineering, Tohoku Imperial University, Sendai, Japan, 1916 to date. Fellow of the Institute.

INSTITUTE ACTIVITIES

May Meeting of the Board of Direction

A meeting of the Board of Direction was held in the office of the Institute on May 2nd. The following were present: L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Donald McNicol, Junior Past-President; W. G. Cady, J. H. Dellinger, R. A. Heising, J. V. L. Hogan, R. H. Marriott, and J. M. Clayton, Secretary.

The following were transferred or elected to higher grades of membership in the Institute: Transferred to the grade of Fellow: C. M. Jansky, Jr.; Transferred to the grade of Member: W. E. Brindley, J. E. Fetzer, C. W. Horn, C. M. Howard, and S. R. Montcalm. Elected to the grade of Member: R. W. Ackerman, A. J. Chesterton, C. R. Hanna, A. H. Morton, O. A. Pearson, and A. M. Trogner.

One hundred and ten Associate members and ten Junior members were elected.

A petition from members residing within the vicinity of New Orleans, Louisiana was submitted to the Board for approval of the formation of a New Orleans Section of the Institute. The petition was approved. Prominent in the organization work of the New Orleans Section were Pendleton E. Lehde, T. G. Deiler, L. J. N. duTreil, and A. Schiele.

Lewis M. Hull was appointed the Institute's representative on the Sectional Committee on Radio of the American Engineering Standards Committee, to succeed Professor L. A. Hazeltine.

1928 Membership Cards

As announced on several occasions in past issues of the Proceedings, membership cards are available for *all* members this year. Cards can be secured only upon specific request to the office of the Secretary, and are not mailed automatically upon payment of member dues.

New Style of I. R. E. Emblem

A new membership emblem, in the form of a screw-back lapel button, approximately one-half the size of the present emblem, will be available within approximately thirty days. This emblem will be finished in the various colors signifying the dif-

ferent grades of membership. It is of 14 k. gold and can be purchased from the Secretary for $2.75 postpaid.

1928 Year Book

The 1928 Year Book of the Institute, containing general information concerning the Institute, the Constitution and By-Laws, a catalog of the membership listed both alphabetically and geographically, and other information of value, is at the printer's and should be mailed as a supplement to either the June or July issue of the Proceedings.

References to Radio Literature

Through the cooperation of the Bureau of Standards, with the July issue the Proceedings will contain the monthly list of references to current radio literature as prepared by the Bureau and printed in the Radio Service Bulletin. The first printing will consist of a recapitulation of all references compiled since the first of 1928, to be followed by the monthly list, in each issue of the Proceedings.

Papers in Pamphlet Form

On page xxxiv of this issue will be found a list of papers which are in pamphlet form and are available for distribution, free of charge, to members of the Institute. Kindly address requests to the Secretary, indicating the papers wanted numerically and *not* by titles. To non-members of the Institute these reprint copies are sold at fifty cents each.

Errata

An error exists in Fig. 2 of the paper "Broadcast Control Operation" printed on page 502 of the April, 1928 issue of the Proceedings. The total resistances of the potentiometers shown should be paralleled and connected to the upper end of their corresponding coil secondaries so that the impedance facing the amplifier input remains sensibly constant, the audio input to any of the potentiometers varying according as the microphone in question is to play a greater or lesser part in the final output.

Material on Formation of Sections

The 1927 and 1928 Committees on Sections have prepared a document outlining the steps to be taken in organizing a Section of the Institute, containing suggestions relative to the successful

operation of a Section and describing many of the important functions of a Section. This material was prepared, primarily, for the information and guidance of members interested in the organization of a Section of the Institute. Members so interested may obtain a copy upon application to the Institute office.

Institute Meetings

New York Meeting

On May 2nd a meeting of the Institute was held in the Engineering Societies Building, 33 West 39th Street, New York. L. E. Whittemore, Vice-President of the Institute, presided.

K. S. Van Dyke presented two papers. The first, "The Piezo-Electric Resonator and Its Equivalent Network," is published in this issue of the PROCEEDINGS.

The second, "Some Experiments with Vibrating Quartz Spheres," is summarized as follows: Quartz spheres, mounted between parallel plate electrodes which may have much greater separation than the diameter of the sphere, respond as piezo-electric resonators to frequencies in the following ratios: 1, 1.47, 1.61, 1.83, 2.12, 2.18, and 4.12. Some of these response frequencies appear as doublets which depend for their relative intensity on the orientation of the sphere in the field. The lowest frequency corresponds to a radio wavelength of about 108 meters per millimeter diameter of the sphere. The decrement of the sphere is very small and the sphere behaves as an oscillator as well as a resonator for some of these modes. The geometry of the modes listed is indicated by the Giebe and Scheibe type of luminous discharge on, and near, the surface of the sphere when it is driven by a sufficiently intense field. Photographs of these discharge patterns were shown for the first three of the above modes for a sphere about 5 cm. in diameter. In strong electric fields in which the sphere is resonant, the sphere becomes very slippery to touch and slides around quite freely, without rolling, on a flat supporting electrode, or slides along a track without rolling, or runs around a ring. Though apparently quite still when resting on a smooth concave electrode, nevertheless, if centered by a small hole, it selects an axis for rotation and rotates steadily about this axis, the rotation being apparently due to the reaction of the sphere on the support rather than to any winds of the Meissner type which it may give off. Winds of the latter type are present, however, but in the

vibration modes observed, apparently have no moment for rotation.

In the discussion which followed these papers, the following, among others, took part: F. K. Vreeland, L. E. Whittemore, K. S. Van Dyke, L. M. Hull, and W. C. Bohn.

It is hoped that the second paper may appear in some future issue of the PROCEEDINGS.

Some three hundred members and guests attended this meeting.

ATLANTA SECTION

A meeting of the Atlanta Section was held on May 9, 1928 in the Civic Room of the Hotel Ansley, Atlanta. Preceding the meeting an informal dinner was held in the Ansley Hotel.

F. H. Schnell, of the Burgess Battery Company, presented a paper, "Aircraft Radio."

Officers to serve until December 31, 1928 were announced as follows: Chairman, Walter Van Nostrand; Vice-Chairman, D. C. Alexander; Secretary-Treasurer, George Llewellyn.

This was the last meeting of the Section until fall.

BUFFALO-NIAGARA SECTION

A joint meeting of the Canadian Section, Rochester Section, and Buffalo-Niagara Section, to which members of the Niagara Frontier Section of the A. I. E. E. were invited, was held on April 11th in Foster Hall, University of Buffalo. L. C. F. Horle, Chairman of the Buffalo-Niagara Section, presided.

J. M. Thompson, of the Ferranti Meter and Transformer Company of Canada, presented a paper, "Characteristics of Output Transformers." The paper dealt with the operating characteristics of the output transformer, which are developed in terms of the known speaker, tube and transformer constants. In the first part of the paper the general formula for the speaker current is developed and the effect of varying the transformer constants shown. The turn-ratio of the transformer for maximum speaker current is considered in relation to the commonly used impedance ratio formula. The limitation of the impedance ratio formula is then pointed out and limits set for its general use. The general form of the current frequency characteristic for exponential horns and dynamic cone speakers is then obtained and a general method for matching the speaker to the output tube is given. In the latter part of the paper, curves are given to

show the results obtained in the mathematical part of the paper. The curves also include the results of tests made in the Toronto laboratory of Ferranti, Ltd., to check the fundamental formula. The effect of the turn-ratio on the form of the current frequency characteristic is shown and a method of using the turn-ratio of the output transformer to match the speaker to the output tube is given. A perfect transformer is also compared with a good commercial transformer and the general effect of the leakage inductance and the self capacity of the transformer was shown.

Messrs. Manson, Million, Horle, Klumb, Hector, and others participated in the discussion which followed.

The second paper of the evening, "Shielding of Radio Receivers" was presented by M. L. Levy, of the Stromberg-Carlson Telephone Manufacturing Company. The paper pointed out the necessity for electrostatic and electromagnetic shielding in radio-frequency amplifiers of high gain. No attempt was made to explain why this shielding was necessary as it was the author's intention to cover practical applications of shielding to receivers of commercial types. With the aid of exceptionally well-prepared slides, Mr. Levy described the fundamentals involved in both electrostatic and electromagnetic shielding and how, in certain applications, combination effects were obtained by the use of the single shields. Then followed a description of several types of commercial receivers employing both types of shielding.

Messrs. Horle, Jones, Manson, Graham, Lidbury, and others participated in the ensuing discussion.

One hundred and seventeen members and guests attended this meeting.

Canadian Section

In the Electrical Building of the University of Toronto a meeting of the Canadian Section was held on April 18th. A. M. Patience presided.

John P. Minton, of White Plains, New York, presented a paper, "Soft Magnetic Materials in Radio." Messrs. Patience, Thomson, Hepburn, Smith, Richardson, Mott, Price, Pipe, and others participated in the discussion which followed.

Forty-eight members and guests attended this meeting.

On May 2nd a meeting of the Canadian Section was held in the Electrical Building of the University of Toronto. A. M. Patience presided. Mr. Clark, of the Bell Telephone Company of Montreal, presented a paper, "Carrier Current Transmission."

Messrs. Patience, Bagly, Soucy, V. G. Smith, A. H. R. Smith, Lowry, Price, and others participated in the discussion.

Fifty-nine members and guests attended this meeting.

In the election of officers for the Canadian Section it was announced that the results were as follows: A. M. Patience, re-elected Chairman; V. G. Smith, Vice-Chairman; C. C. Meredith, Secretary-Treasurer; and J. M. Leslie, Assistant Secretary.

On April 18th a "Ladies' Night Dinner and Concert" was held in conjunction with the Toronto branch of the A. I. E. E.

Cleveland Section

At the May 4th meeting of the Cleveland Section, held in the Case School of Applied Science, Cleveland, Professor John R. Martin presided. Five speakers were presented. H. W. Fay, of the Incandescent Lamp Division of the General Electric Company, gave a review of the recent I. R. E. paper on the UX-250 Power Tube. The essential features of the paper were set forth in a very clear and understandable manner. The reasons for the choice of the various design characteristics were explained. Comparison with the preceding power tubes, 210, 171, 112, and 120 was made.

M. V. King, of the Sterling Manufacturing Company, gave a review of the recent I. R. E. paper on "The Piezo-Electric Resonator and Its Equivalent Network." A striking feature was the simplicity of the equivalent circuit and the extremely low values of capacities involved. The sharpness of resonance obtained from the crystal oscillator was clearly shown from properties of the circuit.

Dean S. Kintner, radio editor of the *Cleveland Plain Dealer*, gave a "Will Rogers" talk on what a radio editor thinks about. Fortunately, a lot of it never gets into print. Some of the thoughts touched upon were the screen-grid tube, television, and broadcast programs.

Kelvin Smith, of the France Manufacturing Company, demonstrated the effect of suppressing certain frequencies, overloaded amplifiers and the meaning of transmission units by means of a new set of records issued by the Bell Telephone Laboratories. Some of the demonstrations bore a striking resemblance to the familiar "sidewalk" radio.

Ralph Worden, radio editor of the *Cleveland News*, gave a critical review of an article "Television Comes to the Home"

which appeared in a recent radio periodical. The old familiar static will soon take on new terrors for the radio-looker of the future. The Kennelly-Heaviside layer will also soon come into prominence due to the appearance of a secondary or reflected image.

Forty-six members and guests attended the meeting.

In the discussions Messrs. Curtis, Catterall, Gimy, and others took part.

Connecticut Valley Section

The Connecticut Valley Section, W. G. Cady, Chairman, presiding, held a meeting on April 26th in the auditorium of the Hartford Electric Light Association. B. J. Thompson, of the General Electric Company, presented a paper, "Characteristics and Uses of Screen-Grid Tubes." The paper centered around approximately fifty lantern slides which gave details of construction of the UX-222 screen-grid tube, as well as characteristics with different screen and plate voltages and with different bias voltages on the control grid. A very detailed account of how and why the tube acts as it does was given. Typical circuit diagrams for radio- and audio-frequency amplification were shown. The speaker mentioned the advantages of this type of tube over the standard three-element type for certain applications in radio reception, and the almost imperative use in the laboratory where circuits are required in which changes in the plate circuit must not affect the grid circuit.

Sixty members and guests attended this meeting.

Los Angeles Section

On April 16th a meeting of the Los Angeles Section was held in Los Angeles. Dr. R. C. Burt, of California Institute of Technology, presented a paper on "The Photo Electric Cell." The paper pointed out, among other things, that the action of a photo electric cell can, sofar, be explained only by the corpuscular theory, while the phenomenon of radio, light, and heat waves requires the wave theory. It is probable that both theories are correct, but the exact relation or connection is not known. The construction of the cell was explained, and some of its uses described. Its uses as a light-operated relay are almost innumerable.

Philadelphia Section

A meeting of the Philadelphia Section was held on April 27th in the Bartol Laboratories of the Franklin Institute. J. C. Van

Horn presided. E. L. Nelson, of the Bell Telephone Laboratories, presented a paper on "Some Recent Developments in Radio Broadcasting Apparatus," which among other things explained in detail the development work on some of the transmitting equipment in the experimental development group of stations located at Whippany, New Jersey.

Messrs. Wilson, Darlington, Frazier and others discussed the paper.

Eighty members and guests attended the meeting.

San Francisco Section

The meeting of the San Francisco Section of March 14th was held in the Club Rooms of the Engineers' Club, preceded by a dinner attended by twenty-five members and guests. L. F. Fuller presided at the meeting. The speaker, A. H. Saxton, operating engineer of the Pacific Division of the National Broadcasting Company, delivered a paper on "The Tendencies of Radio Broadcasting." The paper was centered around the organization and operation of the National Broadcasting Company. The general layout of apparatus used in conjunction with the studios was described, the speaker explaining how different studios could be connected to different channels. The necessity of careful checking of facilities was pointed out. The system used by the Westinghouse Electric and Manufacturing Company in operating two stations on the same wavelength was briefly touched upon.

In the discussion which followed Messrs. Fuller, Lippincott, Brown, and others took part.

Following the meeting the members visited the studios of the National Broadcasting Company in the Hunter-Dulin Building. From thence they proceeded to the control room of the Pacific Telephone and Telegraph Company where, through the courtesy of C. H. Cole of that organization, a description of the apparatus used, together with demonstrations, was provided.

Seattle Section

The Seattle Section held a meeting on March 31st. W. A. Kleist presided. Major H. C. K. Muhlenberg presented a paper on "Radio in Aviation." The paper covered the following points: A review of radio communication between planes and between planes and the ground; relative advantages of telegraph and telephone communication; physical problems of generator location on plane; beacon flying with particular emphasis upon

accuracy, dependability, and distances covered; shielding effect of metal planes.

Messrs. Wilson, Kleist, Williams, Mason, Renfro, and others discussed the paper.

Twenty members of the Section were present.

Committee Work

Sectional Committee on Radio, A. E. S. C.

The following were recently appointed officers of the Sectional Committee on Radio, American Engineering Standards Committee: Alfred N. Goldsmith, Chairman; C. H. Sharp, Vice-Chairman, and Laurens E. Whittemore, Acting Secretary.

Committee on Membership

A meeting of the Committee on Membership was held on May 2nd in the Fraternities Club Grill. The following were present: H. F. Dart, Chairman; I. S. Coggeshall, H. B. Coxhead, and F. R. Brick. The Committee considered a number of plans looking to a continued increase in membership in the Institute throughout the year. Its next meeting will be held on June 6th.

Committee on Admissions

At its meeting held in the office of the Institute on May 2nd, the following members of the Committee on Admissions were present: R. A. Heising, Chairman; Lewis M. Hull, E. R. Shute, F. K. Vreeland. Approximately thirty-five applications for transfer or election to the higher grades of membership in the Institute were considered. The amount of material on hand for consideration by the Committee during the past few months has been such that it has been necessary for the Committee to work a number of hours at each meeting to consider all of the applications.

Committee on Constitution and Law

The first meeting of the newly organized Committee on Constitution and Law was held at dinner at the Fraternities Club on May 2nd. The following were present: R. H. Marriott, Chairman; E. N. Curtis, H. E. Hallborg, and G. W. Pickard. Most of the meeting was occupied in outlining the future work of the Committee. The Committee is proceeding immediately with the revision of the Constitution.

Personal Mention

C. F. Donbar is Chief Engineer of station KQV of Pittsburgh. Mr. Donbar was formerly with the Bell Telephone Company.

J. P. Putnam, recently engineer, radio department of the Martin Copeland Company, is now employed as engineer with the National Company of Malden, Mass.

F. B. Sternberg, formerly director and manager of the Dallas Radio Laboratories is now associated with the P. F. Collier and Son Company in the Newark Branch.

Captain Guy Hill, attached to the office of the Chief Signal Officer, War Department, has been detailed to duty with the Federal Radio Commission as technical advisor.

E. H. Guilford, late Chief Engineer of the Radiore Company of Los Angeles, has become General Manager of the Radiore Company of Canada, Limited, with Headquarters in Montreal.

Captain S. C. Hooper, Head of the Radio Division of the Bureau of Engineering, Navy Department, has been detailed to duty with the Federal Radio Commission in the capacity of technical advisor.

Charles M. Kelly, Jr. has recently become associated with the Electrical Research Products Corporation of New York City. Mr. Kelly was formerly assistant sales manager of the Amrad organization.

O. M. Dunning has recently joined the staff of the Acoustic Products Manufacturing Corporation of Stamford, doing general development work on both receivers and electrical phonograph equipment.

Daniel E. Harnett has recently become associated with the Acoustic Products Manufacturing Corporation of Stamford, Connecticut in the design of radio parts. He was formerly in charge of production of the Murdock Company.

BEAM TRANSMISSION OF ULTRA SHORT WAVES*

By

HIDETSUGU YAGI

(College of Engineering, Tohoku Imperial University, Sendai, Japan)

Summary—Part I of this paper is devoted to a description of various experiments performed at wavelengths below 200 cm. Curves are given to show the effect of the earth and various types of inductively excited antennas called "wave directors." Part I is concluded with a discussion of beam and horizontally polarized radiation.

Part II is devoted chiefly to the magnetron tubes used for the production of very short wavelengths (as low as 12 cm.) and the circuit arrangements employed. It is shown that the geometry of the tube and its external connections are of great importance.

The effect of variation of plate voltage, magnetic field strength and other factors on the high-frequency output, is described.

Introduction

THE general term "short wave" loses much of its lucidness when the range of frequency involved is considered. For this reason, the term "ultra short waves" will apply to only those electro-magnetic waves whose length is less than ten meters.

One of the simplest ways of generating short waves by means of vacuum tubes is to use the push-pull circuit developed by M. Mesny. This connection has been fully described by Mr. Englund in the PROCEEDINGS of the Institute.

Waves shorter than ten meters may be produced with stability, but it is difficult to make ordinary tubes operate satisfactorily below two meters. While electro-magnetic coupling is successfully used in the method referred to above, it seems much better to resort to electrostatic coupling in circuits used for the generation of waves of the length described in this paper. Fig. 1 shows a circuit which has been used in the generation of waves shorter than 100 cm.

Stable oscillations were successfully produced using ordinary tubes in this circuit. Such waves have been utilized to determine the natural frequencies of the various forms of metallic bodies. The characteristics of "wave directors", which will be fully described later in the paper, were thoroughly studied with the short waves produced using this type of generator. However, it was

* Original Manuscript Received by the Institute, January 30, 1928; Revised Manuscript Received by the Institute, March 29, 1928. Presented before meetings of the Institute in New York, Washington and Hartford.

impossible to generate waves shorter than 60 cm. even with this circuit using electrostatic coupling within the tubes.

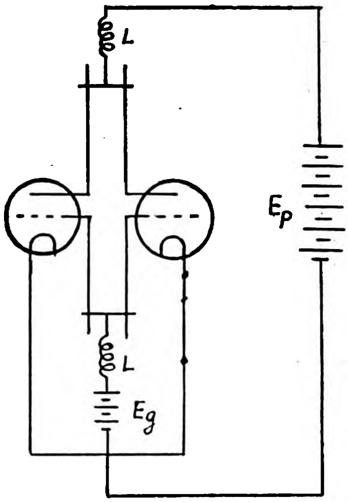

Fig. 1—Circuit Diagram of Oscillator; 60-200 cm.

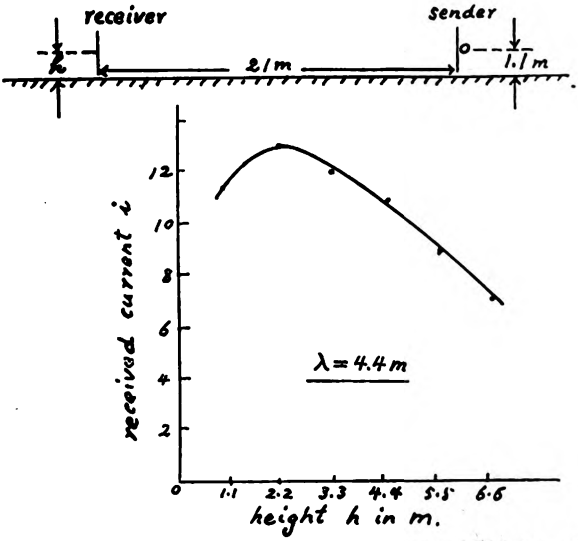

Fig. 2—The Effect of Varying Receiver Antenna Height. Sending Antenna Height Equals 1.1 m.

The method of Barkhausen and Kurz enables one to obtain much shorter waves. By this method, it was possible to reduce the minimum wavelength to 36 cm. using plate voltages in

the order of 300 volts. Schafer and Merzkirch obtained waves of the order of 34 cm. with a plate voltage of 350 volts, and Scheibe has reported a stable minimum of 30 cm. With somewhat less stability, he has produced waves 24 cm. long.

Mr. K. Okabe, assistant professor at the Tohoku Imperial University, has succeeded in generating exceedingly short, sustained waves by introducing certain modifications in the so-called magnetron. These waves are the shortest which it has been

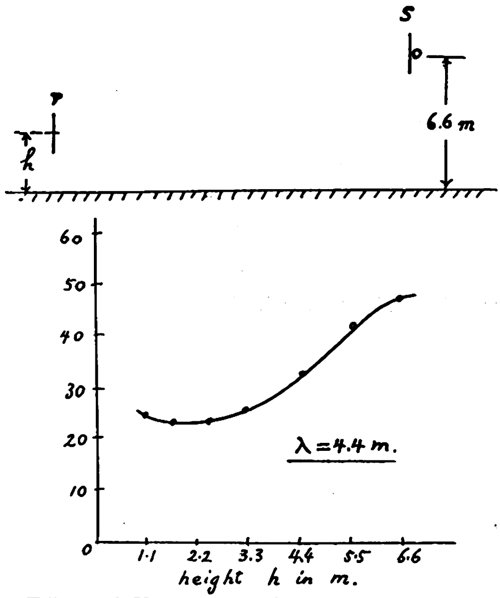

Fig. 3—The Effect of Varying Receiving Antenna Height. Sending Antenna Height Equals 6.6 m.

possible to generate so far as the author is aware. He was able to produce fairly strong radiation at a wavelength of 12 cm. and, by the use of harmonics, was able to obtain a minimum of 8 cm. The practical application of these ultra short waves will be dealt with in Part II of the paper.

Part I

Beam Radiation for 4-Meter Waves

Mr. S. Uda, assistant professor at the Tohoku Imperial University, has published nine papers in the *Journal* of the I.E.E. of Japan on beam radiation at a wavelength of 4.4 meters. Several papers by Mr. Uda and the author have been presented at the Imperial Academy of Japan and the Third Pan-Pacific Science

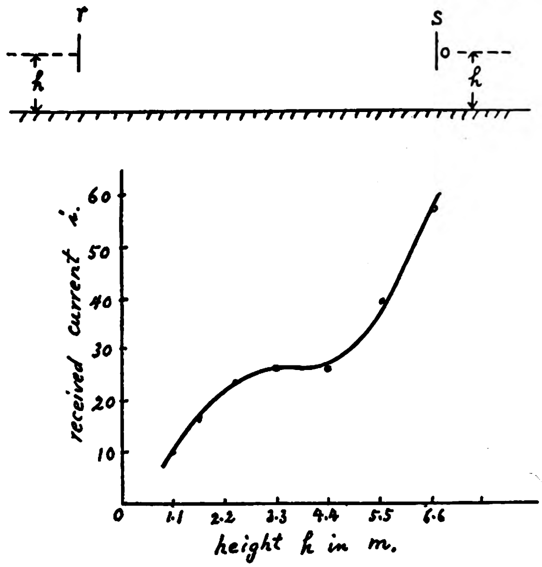

Fig. 4—The Effect of Varying Both Sending and Receiving Antenna Height Simultaneously.

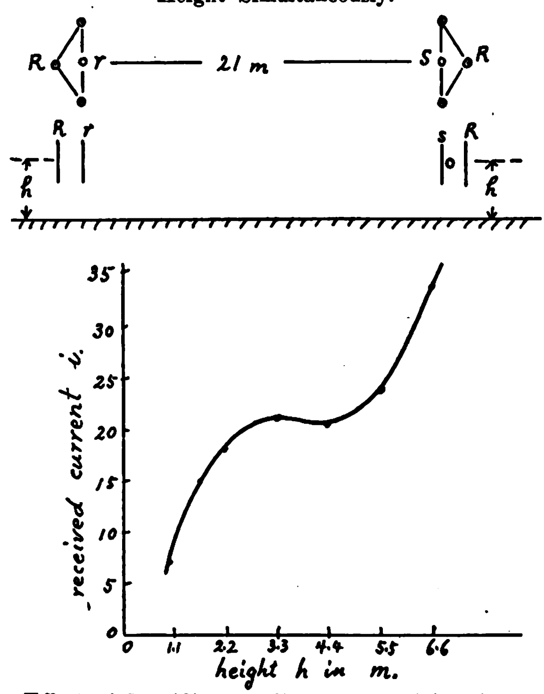

Fig. 5—The Effect of Providing Sending and Receiving Antenna in Fig. 4 with Trigonal Reflectors.

Congress held in Tokyo in 1926. In the following description, some of the much more notable points of the beam system used in this work will be explained. The photographs show some of the actual apparatus used.

Wave Reflectors and Directors

Suppose that a vertical antenna is radiating electro-magnetic waves in all directions. If a straight oscillating system, whether it be a metal rod of finite length or an antenna with capacities at

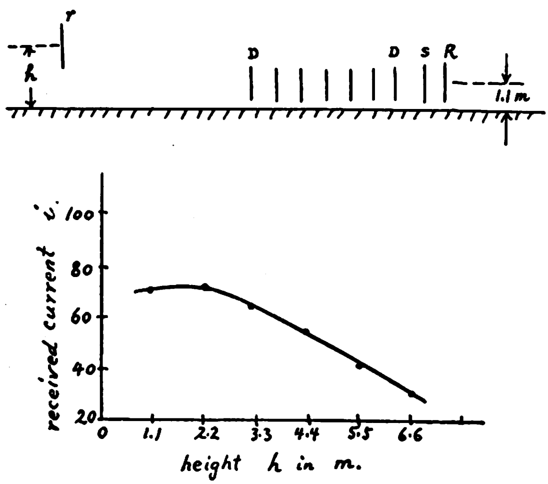

Fig. 6—The Effect of Varying Receiver Antenna Height When Wave Canal Is Applied at Sending Antenna. Sending Antenna Height Equals 1.1 m.

both ends and an inductance at the middle, is erected vertically in the field, the effect of this oscillator upon the wave will be as follows: If its natural frequency is equal to or lower than that of the incident wave, it will act as a "wave reflector." If, on the other hand, its natural frequency is higher than that of the incident wave, it will act as a "wave director." The field will converge upon this antenna, and radiation in a plane normal to it will be augmented. By utilizing this wave-directing quality, a sharp beam may be produced.

A triangle formed of three or five antennas erected behind the main or radiating antenna will act as a reflector. This system will be called a "trigonal reflector." In front of the radiating antenna, a number of wave-directors may be arranged along the line of propagation. By properly adjusting the distance between the wave-directors and their natural frequencies, it is possible to

transmit a larger part of the energy in the wave along the row of directors. Adjustment of the natural frequency of the directors is made by simply changing their length or by adjusting the inductance inserted at the middle of these antennas.

The number of wave-directors has a very marked effect on the sharpness of the beam, the larger number of directors producing the sharper beam. It has been found convenient to designate such a row of directors as a "wave canal."

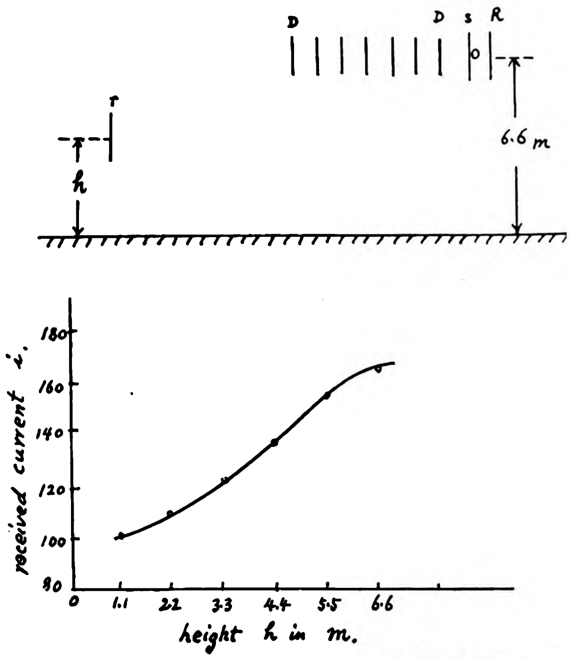

Fig. 7—The Effect of Varying Receiver Antenna Height When Wave Canal Is Applied at Sending Antenna. Sending Antenna Height 6.6 m.

The trigonal reflector and wave canal may also be employed at the receiving station. In this case, the reflector will be called "collector." Here again, the effect of the directors and the wave canal has been found to be considerable.

Radio Beacon

These principles may be used in a radio beacon, by which a beam may be projected in any direction. This is not done by altering the position of the antennas or by revolving the whole system. A number of antennas which are fixed in position are

employed and so arranged that their natural frequencies may be altered between two values. Thus, it may be made either a reflector or a director, depending upon its natural frequency.

The main or radiating antenna is situated at the center, and the others which are used for reflecting or directing the beam are located on two concentric circles whose radii are 1/4 and 1/2

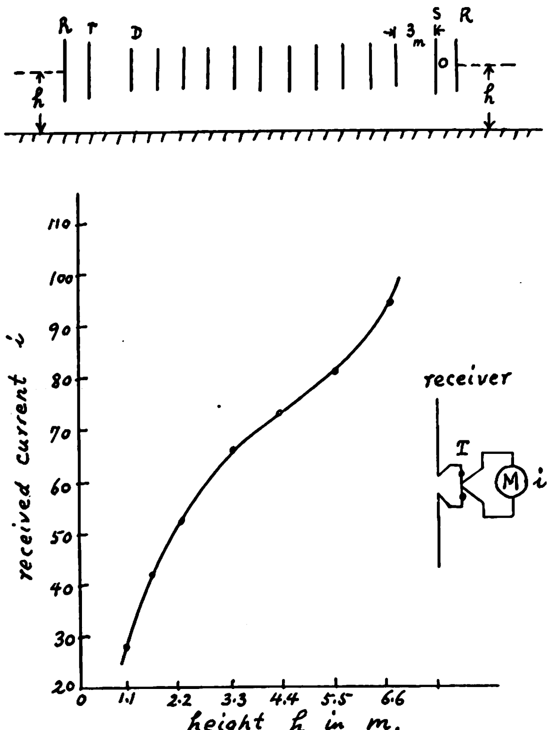

Fig. 8—The Effect of Varying Sending and Receiving Antenna Simultaneously When Wave Canal Is Located between Two Antennas.

wavelength respectively. The direction of radiation may be changed at will by properly controlling the functions of the antennas on these two concentric circles; that is, certain of them are made to act as reflectors while others are made to act as directors of the electro-magnetic wave.

Radio Beacon Transmission

If the sending and the receiving antennas are both surrounded by reflecting systems, and these two structures, which

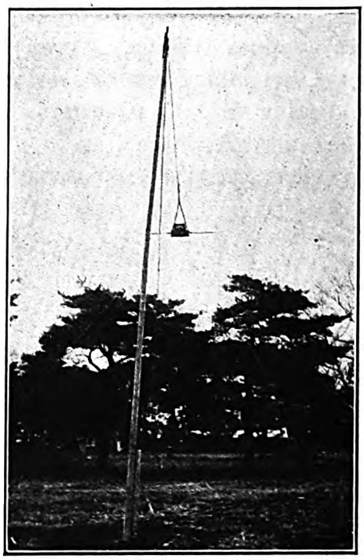

Fig. 9—Horizontally Polarized Wave Receiver in the Air.

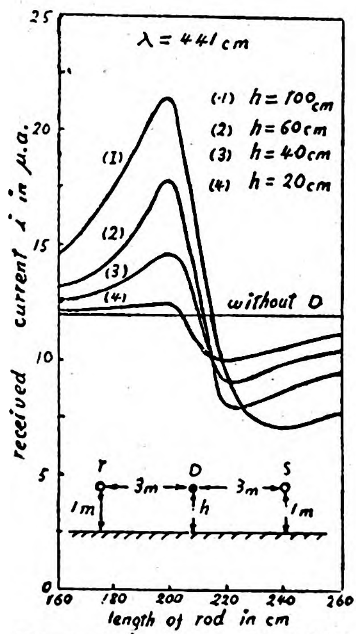

Fig. 10—The Effect of Varying Length and Height of Wave Director on Received Current.

are directed toward one another, are joined by a wave canal, the radio-frequency energy may be directed back and forth along this canal. All the directors forming the canal will have induced oscillations but the intensity and phase displacement will, in

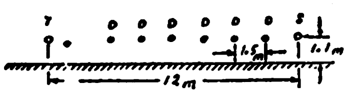

Fig. 11a

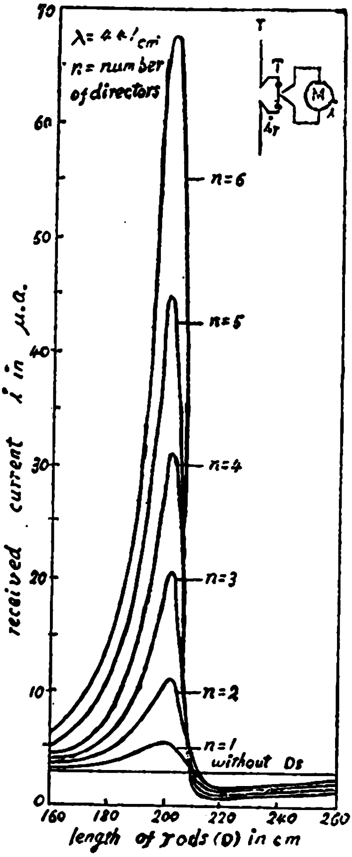

Fig. 11b

The Effect of Varying Number and Length of Directors in Wave Canals on Received Current.

general, be different. A sort of standing wave will exist along the canal and the power will flow at a definite rate from the sending to the receiving station.

The wave energy received can be rectified by means of vacuum tubes or otherwise, and thus it may be used to charge a storage battery. It has been the experience of the author that rectification is very easily obtained even at very short waves.

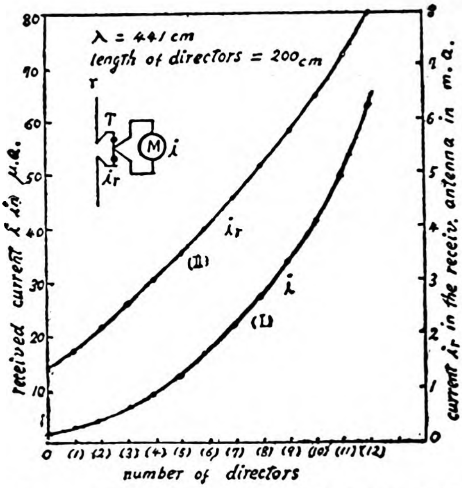

Fig. 12—The Effect of Varying Number of Directors on Received Current.

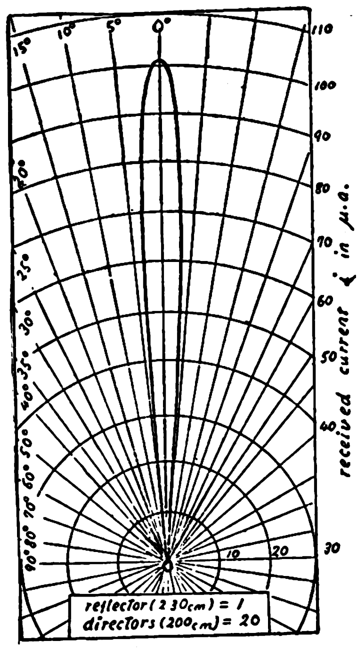

Fig. 13—Beam Radiation from a Radiator Utilizing a Wave Canal.

It appears that the wave collector at the receiving station may suppress to some extent the flow of energy from the sending antenna. It was found that in certain cases it was possible to transmit more power when a certain number of the directors in the middle of the wave canal were removed.

Effect of the Earth

In ultra short wave work, the effect of the earth is very considerable. Some of the experimental results are given below to illustrate this. Figs. 2 to 8, which are self-explanatory, are for various conditions of transmitter and receiver antenna height, with and without trigonal reflectors and wave canals.

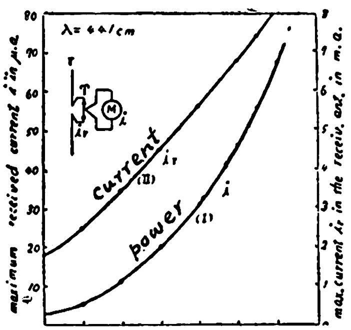

Fig. 14—The Effect of the Number of Directors on Received Current and Power.

It is interesting to note that the energy transmitted increases or is considerably increased when the height of the entire system is increased. As yet, no limit has been found for this effect.

Projector of Horizontally Polarized Waves

A radiating antenna placed horizontally with the earth is naturally directive. The wave is radiated chiefly in a vertical plane bisecting the antenna and perpendicular to it. Various polar diagrams were taken with such an antenna using a receiving antenna such as is shown in the accompanying photograph, Fig. 9. A thermocouple and micro-ammeter located at the middle of this antenna were used to indicate the magnitude of the received power.

The results of these experiments are shown in Figs. 10 to 14. In Fig. 10, *S* and *R* are the sender and receiver respectively, while *D* is a wave director. The effect of varying the length of *D* on received energy is very pronounced, and is a maximum of

about 200 cm., whereas, in the case of vertically polarized waves, this maximum occurred between 190 and 195 cm.

In Fig. 11a, a wave canal is introduced between the sending and receiving antennas and the effect of varying the length of all of the directors is more pronounced than was the case in Fig. 10.

In general, the effect of increasing the number of directors forming the canal is shown in Fig. 12, where i is the current in the indicating meter and i_r is the current in the antenna.

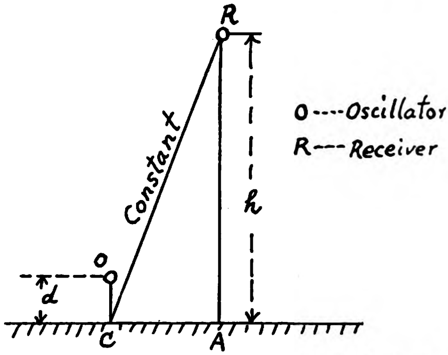

Fig. 15—Diagram Showing the Location of Antenna for Field Strength Measurements.

The length of the directors must be accurately adjusted; otherwise successful directing action will not be obtained. It has been found that the interval between adjacent directors must be adjusted to a suitable value. The most advantageous value for this interval seems to be approximately 3/8 wavelength.

A typical polar curve showing the beam radiation from such a projector is given in Fig. 13. The measurements were taken on a horizontal plane near the earth's surface. Here again, the advantage of utilizing the wave canal at the receiving station is demonstrated to be quite remarkable.

It has been found that power received increases nearly proportional to the square of the number of directors forming the canal. This effect is shown in the experimentally determined curves of Fig. 14.

High-Angle Radiation of Horizontally Polarized Waves

Some experiments were performed in which the field strength around the sending antenna O was measured by a receiving antenna R. The distance CR from R to a point on the surface of

the earth directly beneath the sending antenna was kept constant. The wavelength employed was approximately 260 cm. and the length of the sending antenna O was 135 cm. Fig. 15 shows the arrangement.

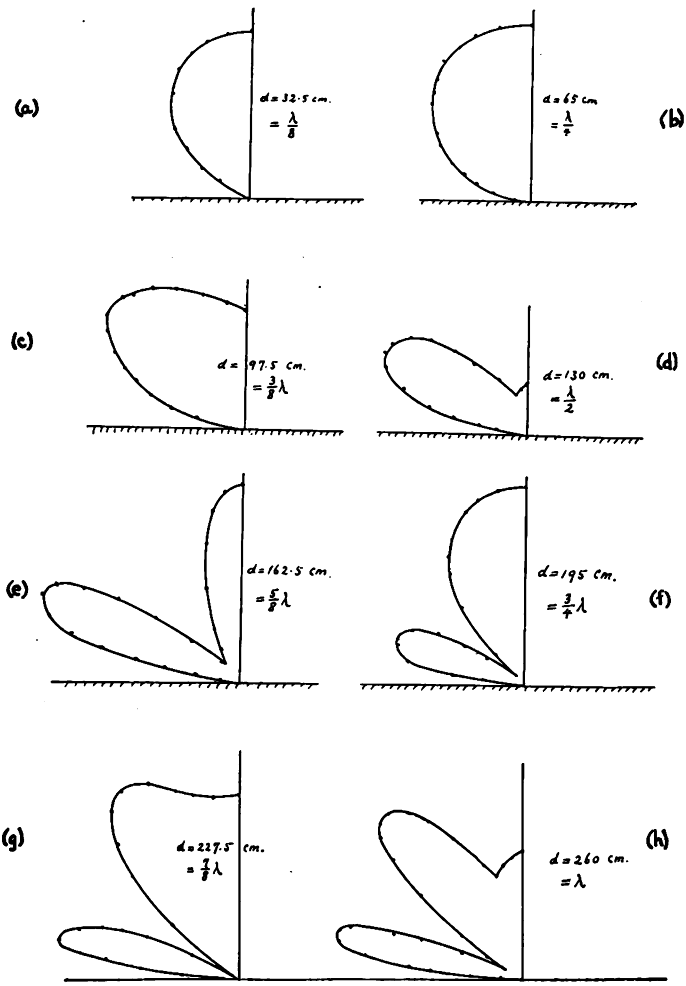

Fig. 16—Polar Diagrams.

The earth seems to act very much like a mirror to ultra-short waves, and reflection from its surface depends upon distance d between antenna and earth as shown in Fig. 15. The experimentally determined polar diagrams shown in Fig. 16, (a), (b), (c), (d), (e), (f), (g), (h), illustrate this fact very well.

The effect of a wave canal upon high-angle radiation of horizontally polarized waves was then studied. A canal was arranged parallel to the surface of the earth in the first case and

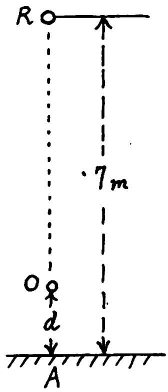

Fig. 17—Diagram Showing Location of Sending and Receiving Antenna for Fig. 18

along the line inclined 30 deg. to the horizontal in the second case. The actual set-up is shown in the two following photographs, Figs. 19 and 20.

It is evident from Fig. 21 that the canal is forcing the beam toward the horizontal direction. Thus, by the use of wave

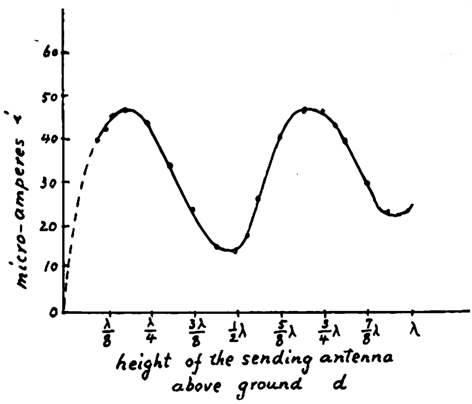

Fig. 18—The Effect of Height of Sending Antenna on Receiver Current.

canals, high-angle radiation may be propagated at various angles to the surface of the earth. This fact may find some practical application in long distance work.

Theory

Theoretical calculations concerning the various experiments described above are naturally involved. Some of the previously mentioned papers presented to the I.E.E. of Japan contain theoretical descriptions of the research. Certain fundamental theories are to be found in a paper which will be published at some later date by the I.E.E. of Japan. This paper will be in English.

Part II

Magnetron Oscillators

A diode is capable of producing oscillations if the anode is a circular cylinder and the cathode is a straight filament at the

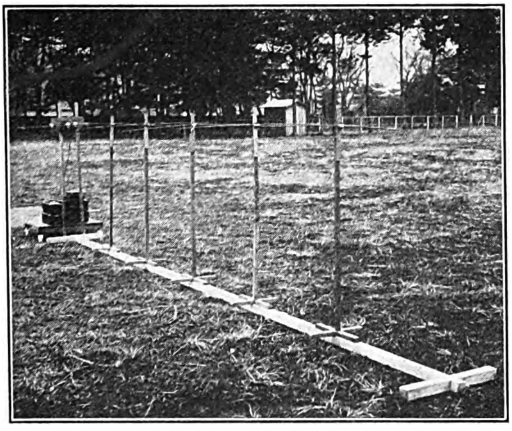

Fig. 19—Projector Horizontally Polarized Wave; 260 cm.

center, with the tube placed in a uniform magnetic field the direction of which coincides with the direction of the axis of the cylinder. When the strength of the magnetic field is increased past a critical value no current should flow through the vacuum tube because the electrons emitted from the filament and attracted by the anode describe circular orbits the diameter of which is less than the radius of the anode. However, when this is tried experimentally sometimes there is residual current flowing to the anode which can be detected by a hot wire instrument. This is evidence of the existence of high-frequency currents.

It has been found that any of the diodes or the triode shown in Fig. 23 can produce short-wave oscillations when sufficiently

high anode voltage is applied and a magnetic field of appropriate intensity is employed. In order, however, that the oscillations be of extremely short wavelength with sufficient intensity, symmetrical construction and exact dimensioning are essential.

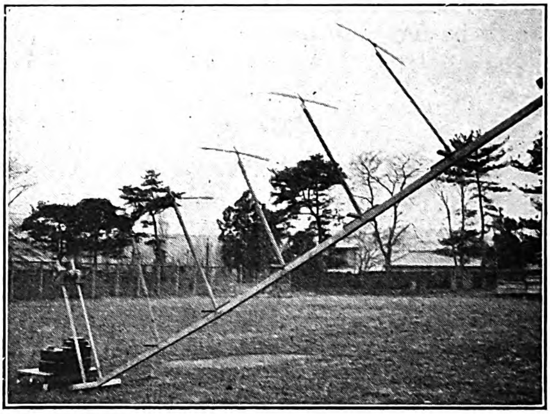

Fig. 20—High Angle Projector of the Horizontally Polarized Wave. 260 cm.

It has been found that the wavelength can be calculated roughly by the following semi-theoretical formula:

$$\lambda_0 = 2\,ct$$

Where
λ_0 = semi-theoretical wavelength
c = velocity of light
t = the time required by an electron for travelling across the space between the cathode and the anode

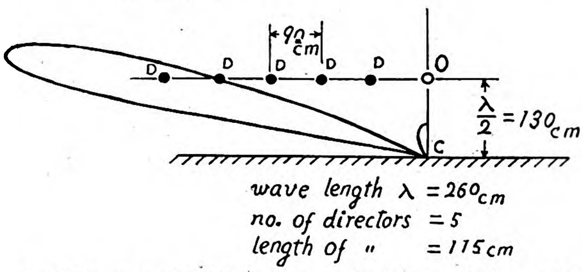

Fig. 21—Polar Diagram with Wave Canal Parallel to Surface of the Earth.

The results are given in the following tabulation. The second column gives the wavelength as measured by the Lecher wire method. The wavelength was practically independent of filament temperature.

MAGNETRON II

D_a=1.32 cm. Vacuum:	L_a=2.5 cm. 10^{-3}/bar.	I_f=3.5 amp. Ni-Anode	l_f=3 cm. W-Fil.
Anode-Voltage (Volts)	**λ (cm.)**	**Intensity of the Oscillations**	**λ_0 (cm.)**
190	150	Weak	87
230	122	"	79
280	88	Middle	72
450	63	Strong	58
500	.....	....	55
1000	32	Very Strong	38
1300	26.5	" "	35
5000	.....	....	17
20000	.....	....	8.5

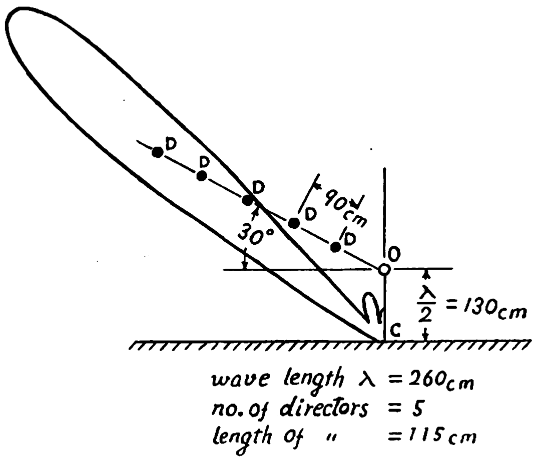

Fig. 22—Polar Diagram with Wave Canal at 30 deg. to Surface of the Earth.

The arrangement of the apparatus is shown in Fig. 24.

The variation of anode current with the magnetic field for a typical tube is shown in Fig. 25. Above the critical magnetic field strength there was still some current flowing which was a

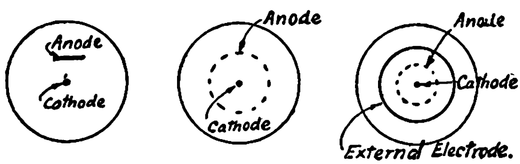

Fig. 23—Types of Diodes and Triode Used Experimentally.

result of the high-frequency oscillations. The most intense oscillation occurred at or near the critical field strength. The oscillations seemed to weaken with increasing magnetization.

When the anode diameter was kept constant larger diameter filaments seemed to give stronger oscillations. More-

over, with larger filament diameters, the anode current most favorable to the production of oscillations was smaller, which is decidedly an advantage.

To get the shortest waves, the anode diameter must be small. The result, however, is that the oscillations become less intense. It was found that the actual length of the anode must not be too short in proportion to its diameter, otherwise the oscillations were very feeble.

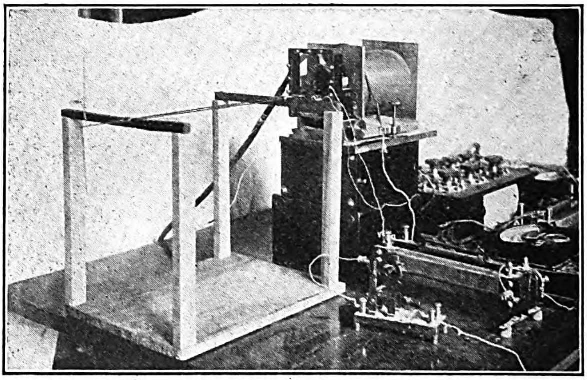

Fig. 24—Apparatus for 26.5-150 cm. Wavelength Production.

The position of the tube in the magnetic field is very important. It was found highly desirable to keep the tube in the most uniform portion of the field. As shown by Fig. 26, a slight deviation from the exact center of the magnetic field coil caused a marked decrease in the oscillation intensity.

Shortest Waves Obtained

Two special tubes of small dimensions were constructed and tried.

No. I	$Da = 4.5$ mm.	$Df = 0.14$ mm.
No. II	$Da = 2.2$ mm.	$Df = 0.07$ mm.

where Da = anode diameter and Df = filament diameter. For the test each tube was placed between the poles of a large electromagnet as shown in Fig. 27.

The relation between the anode voltage and the wavelength for tube No. I is shown in Fig. 28. Tube No. II gave a wave of 19 cm. with 840 volts on the anode and a minimum wavelength of 12 cm. with 1250 volts on the anode. These values of

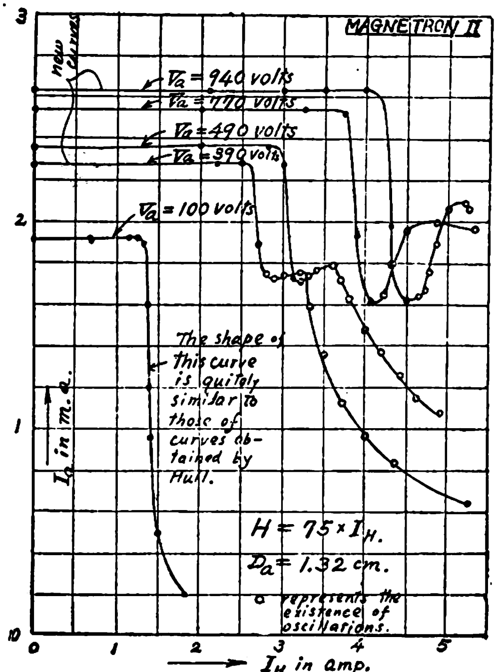

Fig. 25—Variation of Anode Current with Strength of Magnetic Field.

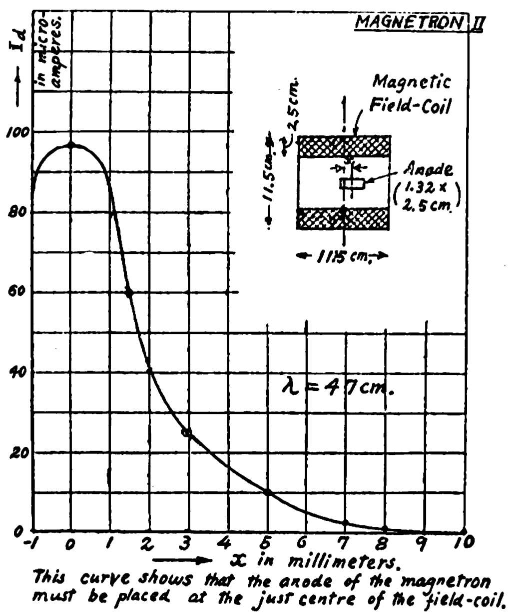

Fig. 26—Variation of Oscillation Intensity with Tube Position in Magnetic Field.

wavelength, however, do not agree very well with the semi-theoretical formula.

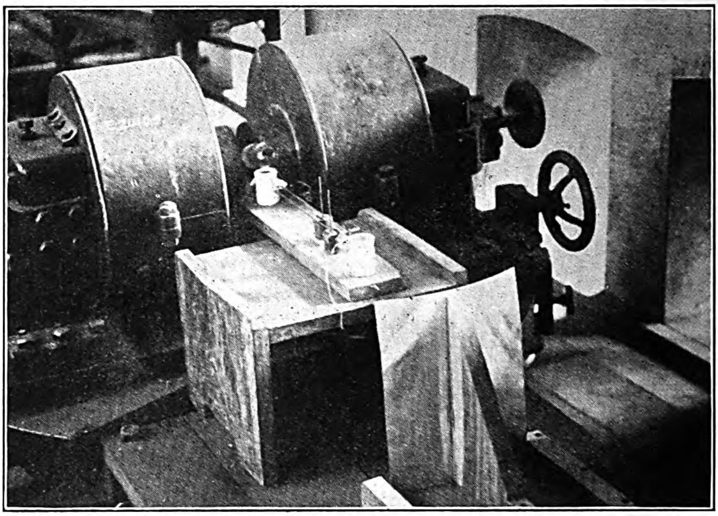

Fig. 27—Apparatus Set-up for the Shortest Waves Obtained (14-15 cm.).

The measurement of the wavelength on Lecher wires was not easy. Too strong a magnetic field seemed to disturb the

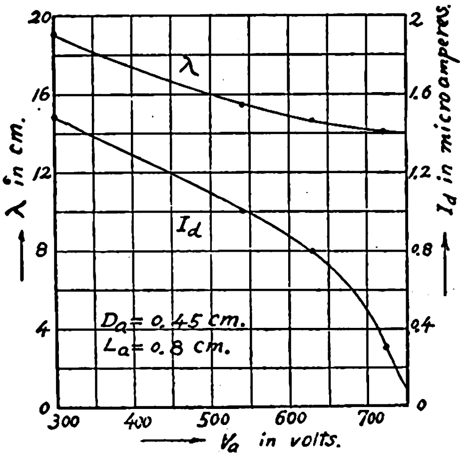

Fig. 28—Variation of Wavelength with Voltage.

steadiness of the oscillations and it was difficult to obtain the shorter waves as a fundamental oscillation. The stronger

magnetic field had a tendency to produce oscillations rich in harmonics.

The most fruitful improvement made was to split up the cylindrical anode into two or more segments by narrow slits cut parallel to the axis of the cylinder. Fig. 29 shows the two-segment type and Fig. 30 the four-segment type of tube. Instead of bringing only one anode lead out of the tube a lead was brought out for each segment. These leads were then

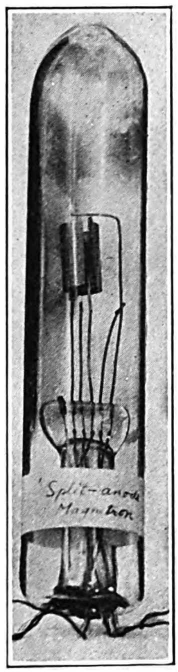

Fig. 29—Split-Anode Magnetron; 2-segment type.

brought together outside of the tube without directly touching each other and brought close to the cathode lead at a point *B* in Fig. 31. After that the leads were all connected and led to the positive terminal of the high-voltage anode battery.

Each anode segment with its leads seems to form a resonant circuit, the natural frequency of which may vary with the length of the lead and the capacity of the segment.

The distance between the anode leads and the cathode lead must also be adjusted at the point *B*, so that maximum oscillation intensity may be obtained. Now, owing to the tuning action of

these resonant circuits, the change of wavelength, due to the change of anode voltage, became inappreciably small. The

Fig. 30—Split-Anode Magnetron; 4-segment type.

wavelength was determined either by a Lecher system or by a receiving set used to indicate standing electromagnetic waves formed before a sheet metal screen.

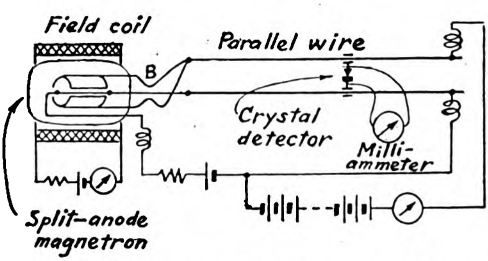

Fig. 31—Circuit Connection for Short Wave Oscillating Magnetron.

40 cm. Waves

A split anode magnetron was found to be especially well suited for the production of very intense oscillations of about 40 cm. wavelength. A typical case is given in the following table.

$Da = 14$ mm. $La = 26$ mm.
$Df = 0.14$ mm. $Lf = 30$ mm.

where La = length of anode and Lf = length of filament.

Anode Voltage	Wavelength	Intensity (Arbitrary)
951	34.5	5.3
724	41.5	15.5
670	42	16
500	42.5	6.7
400	42.5	4
320	42.5	1.8

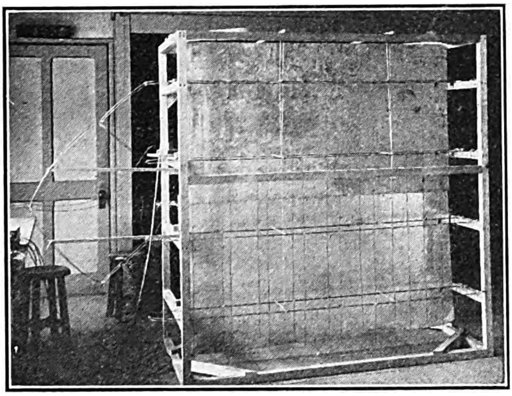

Fig. 32—Antenna System for 40 cm. Wave Transmitter.

The apparatus used in this experiment is shown in Fig. 32 (front view) and Fig. 33 (rear view).

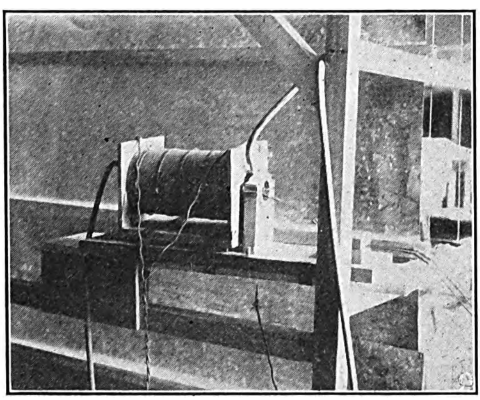

Fig. 33—Magnetron Oscillator for 40 cm. Wave Transmitter.

In order to obtain various directive effects, antenna systems, as shown in Fig. 34, may be used and several of these

may be combined, using metal plates as reflectors; or groups of reflectors as shown in Fig. 35 may be used with parabolic reflectors of sheet metal. Fig. 24 shows a radiating system of

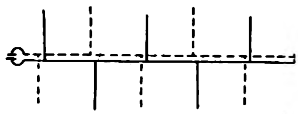

Fig. 34—Directive Antenna, 40 cm.

the type given in Fig. 34. The polar diagrams (Fig. 36) of the antenna system can be calculated from the arrangement of the various elements. The actual measurements showed good

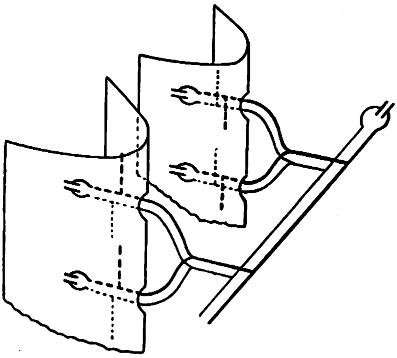

Fig. 35—Directive Antennas.

agreement with the theoretical values and the beam was confined within a small angle in the horizontal and vertical planes.

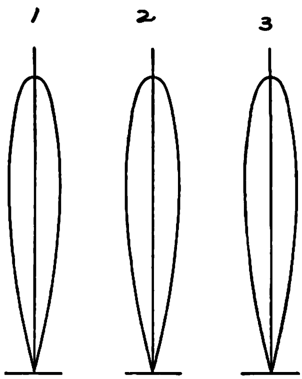

Fig. 36—Polar Curves. 1-Calculation; 2-Horizontal Observed; 3-Vertical Observed.

Reception of Short Waves

For reception a crystal detector or thermocouple attached at the center of a straight antenna can be used. The currents from several such detectors may be combined in parallel or in series, according to the circumstances.

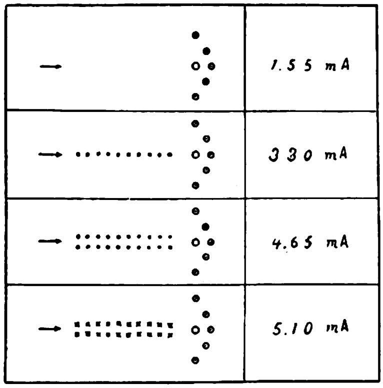

Fig. 37—Reception with a Collector and Wave Canals.

To increase the signal strength a wave collector was built, but its effect was not as remarkable as that of the wave canal. Wave directors on the transmitter proved to be astonishingly advantageous. It was found necessary to use a wave director in order to transmit signals at this short wavelength. The effect of the wave canal at the receiver is shown in Fig. 37.

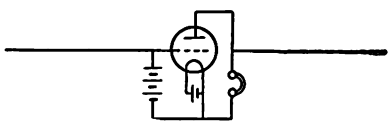

Fig. 38—Receiver (Barkhausen).

The maximum distance covered has so far not exceeded 1 kilometer. In this 1-kilometer experiment the wave was modulated at 900 cycles per second. The exact wavelength was 41 cm. and the anode voltage 1000. A single Hertzian resonator with a crystal detector at the center, a double row of 12-meter director chains and a three-stage amplifier for the modulation frequency were employed. The results are shown in the following table:

1. Single Hertz resonator only	Signal not heard
2. One director chain, no amplifier	Very weak

3. Two director chains, no amplifier	Very weak
4. One chain, with amplifier	Loud and clear
5. Two chains, with amplifier	Loud and clear

The type of receiver suggested by Barkhausen[1] and shown in Fig. 38 was also tried and gave better results than the crystal detector in detecting modulated waves in the neighborhood of $1\frac{1}{2}$ meters.

The experiments described in Part I were made by Mr. S. Uda, and those in Part II by Mr. K. Okabe, to the ingenuity of both of whom the successful development of the beam system is mainly due.

Discussion

J. H. Dellinger:† Professor Yagi's remarkable work stimulates some thought of a radical order. I venture to suggest that before many years radio operations will generally be considered as divided into two classes, broadcasting and directive radio. Radio communication is to a large extent done the wrong way today. And before 1920 radio was all wrong. The only use of radio was for communication between two points, and it was always done by broadcasting in every direction. It was not until 1920 that we had the advent of broadcasting as such, transmission intended for reception by large numbers of receivers. In the eight years since 1920 we have successfully developed broadcasting. At present, therefore, the job of straightening radio out is half done.

It is interesting that 1920 marks not only the rise of broadcasting but also the beginning of directive radio. Ideally, radio transmissions should be broadcast in every direction only when intended for reception in every direction, and should be sent as nearly as possible in one line when intended for reception by one receiver. Since 1920 we have had the gradual and partial evolution of beam systems and other means of confining a communication more or less to the path desired. One instance is the use of a string of relay stations. Now Professor Yagi has shown us that one of the ways to accomplish the directive function is to use a string of absolutely automatic relay stations, viz., the simple devices he calls "directors." Not only in this ingenious suggestion but throughout a wide field of basic possibilities in directive radio,

[1] *Phys. Zeits.* **21**, 1, 1920.
† Chief of Radio Division, Bureau of Standards, Washington, D. C.

Professor Yagi has done exceptional fundamental work and has set forth a series of principles which will unquestionably guide much of the further development. While Professor Yagi's conclusions are validated by experiment, he has, as he says, in many directions only made a beginning and much remains to be done. I am sure that many of those who have heard and those who will read his paper will join him in further pursuit of a number of these interesting possibilities. When they have been fully developed we shall be a long way on the road toward the possibility of carrying on point-to-point communication by directive radio processes.

I have had the privilege of hearing Professor Yagi report not only on the part of his work included in his paper published in the PROCEEDINGS of the Institute of Radio Engineers, but also additional parts of it described in the *Proceedings* of the Third Pan-Pacific Science Congress. His work has included not only this development of wave projectors but also outstanding contributions to the technique of generating and using the shortest of radio waves, the development of the magnetron, and the amusing possibilities of radio power transmission. Whether the use of ultra-short radio waves will be important in long-distance communication, or whether Professor Yagi's ideas will have their principal application in methods of directing radio waves of more usual frequencies, time only can tell. In conclusion, I would like to say that I have never listened to a paper that I felt so sure was destined to be a classic.

THE PIEZO-ELECTRIC RESONATOR AND ITS EQUIVALENT NETWORK*

By

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Summary—*The theory of the piezo-electric and the mechanical behavior of a quartz resonator is stated following Voight and Cady. The functions of the quartz as dielectric and as vibrator are shown to be separable and replaceable by a condenser in parallel with an electrical resonator, i.e., a series chain of inductance, resistance, and capacity. For a Curie-cut quartz rod excited lengthwise through the transverse piezo-electric effect into compressional vibration at the fundamental frequency the series elements become* $L = M/4\epsilon^2 b^2$; $R = N/4\epsilon^2 b^2$; $C = 4b^2\epsilon^2/g$. *This mode of vibration may be termed the "fundamental normal mode" since the vibration direction is normal to the electric field. For a Curie-cut quartz plate excited through the longitudinal piezo-electric effect into compressional vibration at the fundamental thickness frequency, the series elements become* $L = e^2 M/4\epsilon^2 l^2 b^2$; $R = e^2 N/4\epsilon^2 l^2 b^2$; $C = 4\epsilon^2 l^2 b^2/e^2 g$. *This mode of vibration may be termed the "fundamental parallel mode" since the vibration direction is parallel to the electric field. M, N, and g are respectively the half-mass, mechanical resistance factor, and "equivalent stiffness" of the rod or plate whose thickness along the field is e, and dimensions normal to the field are l and b, the latter being along the optic axis. The parallel capacity is shown to be less than that for a quartz dielectric which is free to assume piezo-electric strain and to be equal to that for unstrained quartz. Phase and amplitude variations of current to the resonator are shown as obtained with the cathode ray oscillograph.*

HALF a century ago the Curies discovered the piezo-electric effect. In the *direct effect* charges result from the compression of a crystal, which reverse in sign if the crystal is stretched instead of compressed. There are other types of stress under which similar charges appear and reverse in sign with the reversal of the stress. In the *converse effect,* which Lippmann showed must exist whenever a direct effect is thus reversible in sign, strains result when potential differences are applied to the crystal. These strains also reverse in sense with changes in the sign of the P.D.

The feebleness of these effects may be seen from the following data on a quartz plate, 1 mm. thick and 2.5 cm. square, the thickness lying along some optimum direction in the quartz for electrical effects from pressure. This direction is called *an electric axis* of the crystalline material, and *the electric axis* for a plate

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cut from the quartz as above stated. It was with plates cut in this fashion that the Curies measured the charges which result from pressure. A weight of 1 kilogram on the surface 2.5 cm. square should reduce the thickness by about two parts in ten million and set up in addition to this elastic strain a dielectric strain as well. This is the familiar Maxwell electric displacement due to polarization of the dielectric. With electrodes in contact with the square faces of the plate and connected to an external condenser of equal capacity the P.D. between the plates would be about 1 volt instead of 100.

The P.D. which results would be smaller as the condenser has larger capacity. In the converse effect, if a sufficient charge were applied to the electrodes to cause a P.D. of 100 volts the quartz plates would become thicker or thinner by two ten millionths of a millimeter, i.e., about one half of one thousandth of a light wave, if the plate were free to change its thickness. If not free, it would exert a force of a kilogram on its confines.

When a P.D. applied along the electric axis of a quartz plate tends to cause the quartz to contract along this axis, there is another direction at right angles to this electric axis along which the plate tends simultaneously to expand. Fractional changes in length along this second direction *transverse* to the applied field are equal to those which occur along (*longitudinal* to) the electric axis and are likewise reversible in direction with the sign of the field. It is with these transverse effects that Cady's[1] original development of the piezo-electric resonator was primarily concerned. Resonators cut from the quartz so as to utilize this transverse effect are used chiefly for the longer wavelengths. The two directions which have been referred to as *longitudinal* and *transverse* with respect to an electric axis determine a plane which is perpendicular to the well-known *optic axis* of quartz.

The illustrative examples given have been computed from formulas given by Voigt in his "Lehrbuch der Kristallphysik," formulas based on a theory which satisfactorily predicts the known facts of piezo-electrictification. In view of the minute effects described it is little wonder that piezo-electric phenomena were for so long of little interest except in connection with crystal theory. That so complete a theory of piezo-electricity as that of Voigt should have been developed in a field of so little apparent

[1] Proc. I.R.E. 10, 83, 1922.

practical usefulness is in itself a monument to theoretical science.

Given the phenomena of resonance in a piezo-electric crystal, the effects are no longer so small as to be unimportant. Since its introduction by Cady in 1921, interest in the *piezo-electric resonator* and its applications has developed rapidly until it is at the present day an essential element of most constant-frequency devices in the radio field.

The phenomena of resonance are fundamentally the same whether in a mechanical or an electrical system. A series electrical circuit is said to be in resonance with an applied alternating e.m.f. when the current is in phase with the e.m.f. The current may then be much larger for the same applied e.m.f. than at other frequencies and is given directly by Ohm's law. The natural frequency of free damped oscillations in the circuit is slightly lower than the resonance frequency by an amount depending on the decrement. When an alternating driving force has a frequency nearly the same as the natural free vibration frequency of the mechanical system being driven, then the resulting mechanical displacement may be many times larger than for a steady force of equal magnitude. Resonance strictly occurs for that driving frequency at which the mechanical vibrator so adjusts its response that its velocity is in phase with the driving force. The driving force is then applied directly against the resistance to the motion. For frequencies lower than this the velocity leads the driving force and the impedance which the mechanical vibrator offers to the force is said to have a negative angle. For frequencies above resonance the velocity lags behind the force and the mechanical impedance has a positive angle. The increase in amplitude at resonance is greater as the vibrating system has less damping.

In the quartz resonator the alternating mechanical forces which drive the mechanical resonator result, through the converse piezo-electric effect, from the alternating P.D. applied across its electrodes and are properly timed for resonance when the applied alternating potential has a frequency practically equal[2] to a natural free vibration frequency of the mechanical quartz resonator.

[2] Very slightly higher than the frequency of free damped vibrations. That this is so slight is due to the extremely low decrement of these resonators as compared with decrements met in electric circuits.

We shall see that the velocity of the vibrating crystal at resonance is practically in phase with the P.D. across its electrodes. This will suggest that the vibrating crystal responds like a tuned electric circuit with its current somehow masquerading as a mechanical velocity. Those who are familiar with mechanical impedance will note that the piezo-electric nature of quartz merely offers a coupling link between the mechanical impedance of the vibrator and current in the circuit in which the crystal is used.

Quartz resonators have such low damping that even when vibrating in air the increase in amplitude at resonance may be several thousand-fold. Cady found a 4000-fold increase for a typical plate even when damped by the presence of electrodes as close as possible without contact. If the 1 mm. thick quartz plate, which we have used in illustrating the feebleness of the static effect, should increase its amplitude a thousand times at resonance, an applied resonant P.D. of 100 volts across its electrodes would give it an amplitude of about half the wavelength of visible light. But to have attained a similar degree of compression from a steady applied force a weight of a ton would have been necessary, if the crystal did not break under the strain.

Cady in his original development of the theory of the resonator used the more or less familiar admittance circle diagram for studying the effects of the crystal at frequencies near resonance. The application of the resonance circle offers many advantages to those familiar with its use, as is true of most vector methods of solving alternating current problems. Some electrical engineers who are used to recognize circuit behavior from the arrangement of the elements in a circuit may understand the behavior of the resonator more easily if it can be replaced by a group of ordinary circuit elements. One purpose of the present paper is to develop a network which can, on paper at least, replace the crystal in all of its effects on the circuit in which it is placed. There is a certain fundamental reasonableness in the use of such an equivalent network since the resonator is used in electric circuits, where design problems involve relations between electric circuit elements only. Current flowing in a circuit can know only the effects of capacity, inductance, resistance, and electromotive force and their variations, and the current adjusts itself in frequency, magnitude, and phase to these elements.

For a resonator having a single degree of *mechanical* freedom

the equivalent network is a simple one, in order that it may have a single degree of *electrical* freedom. The actual piezo-electric resonator under consideration is a rod or a plate of quartz which has a great many vibrating modes in which it may show resonance phenomena. Each of these modes may be considered by itself, and simpler mechanical resonators devised, each having a single degree of freedom, and each reproducing the mechanical vibration of the crystal in the vicinity of the resonant mode for which it was devised. The present paper will translate two of these simpler mechanical resonators into capacitance, inductance, and resistance in the crystal circuit. Corresponding to every other resonant frequency of the mechanical piece of quartz there is a similar network of constant elements which can completely represent the quartz in its effect on the circuit at all frequencies except in the neighborhood of other resonant frequencies of the crystal.

The two modes of crystal vibration with which the present paper deals are those lying in the plane perpendicular to the optic axis with end motions of the plate either at right angles to or along the electric axis of the plate and the applied field. Unfortunately there is some confusion between the terminology used by different writers arising from the use of "longitudinal," by some for compressional vibrations the long way of the plate, and by others for compressional vibrations which are set up in the thickness of the plate by the longitudinal piezo-electric effect. These latter call the vibration resulting from compressional strains of the transverse piezo-electric effect a "transverse" mode. Each of these modes of vibration represents the stationary wave pattern of a system of compressional waves (in mechanics and acoustics said to involve *longitudinal* vibrations), and neither vibration is concerned with waves which are transverse in the mechanical sense.

To avoid the confusion due to this double use of the term longitudinal to characterize both compressional vibrations and the piezo-electric effect which may cause the vibrations, the present paper will call modes where the vibration is parallel to the electric axis and the field *parallel modes* and will use the term *normal modes* where the vibration is perpendicular to the electric axis, i.e., at right angles to the field. The parallel modes are most directly driven by the longitudinal piezo-electric effect, and the normal modes by the transverse piezo-electric effect. The trans-

verse and longitudinal piezo-effects may combine through elastic reactions in the quartz to drive the plate in a state of resonance for compressional waves along the third dimension of the plate, which is here the optic axis. Resonance frequencies for such vibrations are found by Hund.[3] To these *optic axis modes* the results of the analysis in the present paper are not directly applicable, though the method can be extended to apply to them.

Fundamental modes are those for which the wavelengths of these compressional waves are in the parallel case twice the thickness, and in the normal case twice the length, of the resonator. In either case the middle of the quartz is considered as a node with the ends in greatest motion. Higher overtones are not strictly in harmonic relation to these fundamental modes and have two or more nodal planes through the crystal. The networks to be derived are those for the parallel and normal fundamental modes of vibration of quartz plates or rods.

There are still other fundamental modes of vibration of quartz resonators, one of which is in common use. This Cady[4] has shown to be a shear mode. To excite it requires that the plate shall be cut from the original crystal with its face parallel to one of the six prismatic faces of the quartz instead of perpendicular to this face as for the plates with which this paper deals. In the plates with which the present paper is directly concerned, often described as "Curie Cut" plates, the electric field is along a natural electric axis of the quartz. Plates cut so as to use the shear are frequently referred to as "30 deg. crystals," as in their use the field makes an angle of 30 deg. with an electric axis. These latter are preferred by some for broadcasting station control. Their natural radio wavelength corresponds to about 140 meters per mm. thickness of the quartz as against about 110 meters per mm. with "Curie Cut" crystals. Though the analysis of the present paper is applicable only to "Curie Cut" crystals, a similar method may be applied to the shear vibrations of these "30 deg. crystals." Indeed it may be applied to any type of mechanical vibration which produces through known laws a change in an electric current, provided this change reverses in direction with the direction of the mechanical effect, and a network may be derived to have identical effects on the current.[5]

[3] Proc. I.R.E., 14, 459, 1926.
[4] *Phys. Review*, 29, 617, 1927.
[5] The general theory of such equivalent networks is given by Butterworth, *Proc.* Physical Society, 26, 264, 1914, and 27, 410, 1915.

The following familiar electrostatic theory[6] is assumed. Consider two parallel plates each of area A and separation e with an intervening medium of dielectric constant k. This dielectric may be said to have electric susceptibility n such that $k=1+4\pi n$. This is analogous to the $\mu=1+4\pi k$ relating magnetic permeability to magnetic susceptibility. Opposite charges of surface density σ exists on the inner surface of each plate. If the P.D. between the plates is V the electric intensity, or strength of the field, between the plates is V/e. The electric displacement in the dielectric is $P=kV/4\pi e$ in units of Faraday tube per sq. cm., and is also equal to σ. The polarization P is nV/e.

Voigt shows that a polarization P may result in quartz either from an applied external mechanical pressure p normal to proper faces of the plate, from mechanical shear in the quartz, and from an applied external field V/e. With shear and its associated polarization the present paper is not concerned. The polarization from pressure depends on the magnitude of the stray capacity of the closefitting electrodes and on any other capacity to which the electrodes are connected. Tubes of polarization in the quartz terminate on the inner surfaces of the electrodes resulting in charges of opposite sign on the two electrodes. These charges produce in the quartz dielectric between the electrodes their own polarization in a direction opposite to that from the pressure. The resultant of these two polarizations determines the force field between the charges on the inner surfaces of the electrodes and the force field determines the P.D. between the electrodes.

Furthermore a field applied from an external source produces a polarization in the quartz. A part of this polarization gives the usual dielectric displacement of a quartz condenser. An additional part results in a mechanical strain in the quartz, or else in a stress to balance some externally applied mechanical stress.

The properties of quartz which are responsible for these effects are given by the piezo-electric constant and the piezo-electric modulus, designated by ϵ and δ respectively. These should, strictly speaking, have subscripts to denote the directions to which they relate in the quartz. In this paper, the former, ϵ, is the polarization along the electric axis resulting from unit mechanical compressional strains s in the quartz. This associated strain may be either longitudinal or transverse to the electric

[6] All electric quantities used in this paper are expressed in c.g.s. electrostatic units unless specifically stated otherwise. Mechanical quantities are in c.g.s. units.

axis, but is always in the plane perpendicular to the optic axis. The latter, δ, is the polarization along the electric axis associated with unit mechanical stress along either of these same two directions. The values of ϵ and δ are -4.77×10^4 and -6.45×10^{-8} respectively in c.g.s. electrostatic units (Voigt's values). The quartz crystals is, of course, extremely complex in its elastic and its electrical behavior, showing different properties in different directions. Because of the very involved, and in fact unknown, exact nature of the strains involved in the vibrating rods and plates with which this paper is to deal, the rigorous equations which Voigt gives have been simplified to the extent of replacing algebraic sums of various component ϵ's and δ's by the single term most significant for our problem. For the complete expressions from which equations (1) to (4) are taken reference should be made to Voigt.

Voigt expresses the facts of piezo-electric polarization as they concern us here by the equations:[7]

$$P=(n+\epsilon\delta)V/e-\delta p=n'V/e-\delta p \tag{1}$$

$$\sigma=V(1+4\pi n')/4\pi e-\delta p=k'V/4\pi e-\delta p \tag{2}$$

$$Q=C'V-A\delta p=C'V-A\epsilon s \tag{3}$$

$$s=\delta V/e-up;\ p=\epsilon V/e \tag{4}$$

The two parts of the polarization are seen in (1). The first part involves the susceptibility n' of a piezo-dielectric. As is seen in this equation the dielectric has susceptibility from two causes, n which any ordinary dielectric would have, and $\epsilon\delta$ which is peculiar to its piezo-electric nature and accompanies its expansion or contraction due to the filed under constant end pressure. The effect of this latter is to increase slightly the apparent capacity of a condenser having a piezo-dielectric above that which the identical condenser would have if its dielectric were to be somehow robbed of its piezo-electric property, or if the crystal could be prevented from straining under polarization. This increase of capacity when extension at constant stress is permitted is of the order of one per cent and is probably of theoretical interest only; furthermore, ordinary measurement of the dielectric constant of quartz yields the *apparent* value $k'=1+4\pi n'=1+4\pi(n+\delta\epsilon)$. Hereafter the condenser functioning through this apparent dielectric constant will be referred

[7] Adapted from "Lehrbuch der Kristallphysik," 915–919.

to as a *piezo-condenser,* to distinguish it from the *piezo-vibrator* which is simultaneously involved in the same piece of quartz.

The second part of the polarization in the quartz is δp of equation (1). This accompanies a compressional strain s in the quartz, or, through the strain, its associated stress p. ($\delta p=\epsilon s$). The ratio $\epsilon/\delta=u$ is an elastic constant of the quartz and is roughly equivalent to Young's modulus in the plane perpendicular to the optic axis, in view of our approximations. Equations (2) and (3) express the density σ of the charge on the inner surfaces of the electrodes, and the total charge Q, when the area of the electrodes is A. C' is the measurable static capacity of the piezo-condenser when the crystal is under a constant stress. Equations (4) are fundamental equations for stress and strain in a piezo-electric substance, from which equations (1) to (3) are derived.

Our working formulas (3′) for the P.D. across a piezo-electric resonator come from a rearrangement of (3) above with allowance for additional capacity C'' which may exist in parallel with the resonator electrodes. C'' may perhaps represent a leakage capacity from the outer surfaces of the electrodes and the leads, or an extension of the electrodes beyond the area A of the crystal face. Also a modified strain s',[8] which is constant through the quartz, is used to avoid the actual strain distribution through the quartz.

$$V=\frac{(\sigma+\delta p)A}{C'+C''}=\frac{(\sigma+\epsilon s')A}{C'+C''}=\frac{Q+A\delta p}{C'+C''}=\frac{Q+A\epsilon s'}{C'+C''} \qquad (3')$$

The illustrative values already given to show the feebleness of the static effect were obtained by substitution in these formulas. The additional parallel capacity C'' may be considered to be included in C'.

When the mechanical quartz plate is vibrating in resonance with an applied P.D. it has been stated that strains may exist which are a thousand times as large as for the same static P.D. It is evident from (3′) that in static strains $\epsilon s'A$ now plays a part similar to that which charge plays in ordinary condensers. If, on the other hand, in considering vibratory phenomena, and thus, alternating strains, we regard s' as the r.m.s. value of the strain, that part of the incident current which is responsible for resonant strains is $2\pi f_1\epsilon s'A$ (effective or r.m.s. current) if f_1 is the frequency

[8] The modified strain s' is explained later, and it is there shown that it may be used in place of s in applying Eqs. (1–4) to the resonator.

at resonance. This amounts to a resonant current of 10 milliamperes or so to the $2.5\times2.5\times0.1$ cm. plate previously used in the static illustrations. It is here considered to be driven in resonance by a 100-volt 2800-kc. source.

The charging current to this same resonator due to its piezodielectric properties alone is an added 3 microamperes which flows even when the resonator is damped so that it cannot resonate. These two current values are computed from expressions $2\pi f_1 \epsilon s' A$ and $2\pi f_1 V C'$.

Thus it is seen that it is the actual straining of the quartz in its resonant mechanical vibrations that is responsible for most of the current to the resonator. This straining, and hence the current, assumes a given value for smaller applied P.D.'s at resonance than away from resonance, a fact which is also characteristic of currents in series resonant circuits. Current to the mounted crystal electrodes whether at resonance or not goes into two parallel channels, one channel that of the condenser made of piezo-dielectric and any parallel condensers, and the other channel that of the vibrating crystal. The relative magnitude of these at a given frequency and for given crystal area depends only upon the capacity of the condenser and the actual magnitude of the vibratory motions which the given crystal performs. This latter is limited by the mechanical impedance of the vibrator and by the driving P.D. This statement is analogous to the ordinary law for alternating currents in a divided circuit. For the divided circuit network, which is to be derived as equivalent to the piezo-electric resonator so far as circuit phenomena are concerned, a corresponding statement is: At any frequency, whether at resonance of one of the branches or not, the alternating current I divides into two channels with relative magnitudes which depend only on the impedances of the two branches.

Caution is needed in applying equations (3′) to the vibrating crystal. These equations hold for instantaneous values of the P.D., charge, pressure, and strain. With sinusoidal variation of these quantities they must be used in proper phase relation, e.i., they must be treated as vector quantities. These same equations are true if vector quantities are used. The problem is simplified by using the time derivatives of this equation, made explicit in current, Eq. (6). This total current to the resonator and its associated capacities will then be seen to be the vector sum of

that to the piezo-condenser and that to the piezo-vibrator. This vector sum, or a corresponding algebraic sum for instantaneous values of current, corresponds to Kirchoff's law for current at a junction. Thus the current to the vibrating crystal and its associated condensers behaves as if it flowed to a condenser and some other circuit in parallel. The problem is to find the nature of this parallel branch.

Before proceeding to the derivation of the network it is necessary to recall Cady's treatment of the theory of longitudinal vibrations of viscous rods. In this, as has been suggested, he shows that when lateral effects are neglected and compressional waves are travelling along the rod driving it at a frequency near the fundamental frequency, the analysis of the vibration may be simplified by substituting for the actual rod an equivalent vibrating system possessing a single degree of freedom. This equivalent system is so designed as to have an amplitude of vibration equal to that of one end of the actual rod (whose middle is fixed) when the driving force for the mechanical substitute system is equal to twice the force on one end (the sum of that on two ends) of the actual vibrating bar. His equation of motion is:

$$F_0 \cos \omega t = M\ddot{x} + N\dot{x} + gx \tag{5}$$

Here F_0 is the maximum amplitude of the sinusoidal force applied to the equivalent system. ω is 2π times the frequency of the applied force; M is half the mass of the rod; N is the mechanical resistance coefficient, which is equal to $\pi^2 \rho A' Q/2l'$, and Q is the viscosity of the resonator material including any viscous effects of the atmosphere and the mounting in which it vibrates. ρ, A', and l' are respectively density, vibrating face area, and length of the rod. g is the *equivalent stiffness* for the substitute system and is related to Young's modulus G for the actual rod through the equation $g = \pi^2 A'G/2$. x is the instantaneous displacement of one end of the rod. Dotted letters are used for first and second time derivatives. Cady also shows that the fictitious instantaneous pressure which may be supposed to act on each end of the vibrating bar to drive it in resonance is the same as the uniform real stress acting throughout the bar.

Most of the analysis which has thus far been presented is an adaptation from Voigt[9] or from Cady.[10] Cady proceeds to find the current to the resonator and its mounting both analytically

[9] Voigt, "Kristallphysik."
[10] Cady, *Phys. Rev.* 19, 1, 1919; Proc. I.R.E., 10, 83, 1922.

and graphically with the aid of the resonance circle. We shall, however, further interpret our equation (3′) for potential difference across the electrodes of the resonator. Differentiating (3) and solving for I gives equation (6) the current to the resonator.

$$I = \dot{V}C' - \epsilon A\dot{s}' \tag{6}$$

In the mechanical effects with which this paper is concerned, those of rods vibrating near resonance, the strain s obviously varies from point to point in the rod at any instant. A stress, which may be thought of as causing this strain at each point, varies similarly from point to point. Since this stress is p of Eq. (1), that component of the polarization P which is associated with this stress obviously varies according to the same law along the rod as does the strain. The corresponding component (by Eq. (3)) of the total charge on the electrodes is $2l'$ times the integral of this polarization over half the length of the plate. Since the instantaneous extension of the rod is the same function of strain through the half length of the rod, we can replace the point strain in Eq. (3) by a modified strain s' which is uniform along the rod, but varying with time and which is defined by $s' = 2x/l'$. Here x is the actual mechanical displacement of the end of the rod and s' simply the average strain through the rod. In effect it replaces the strain of the actual rod by that for a mechanical resonator of a single degree of freedom.

This "average strain" is made up of two components at each instant in time. One part is that which accompanies the $\epsilon\delta V/e$ part of the polarization in Eq. (1) by virtue of the piezo-dielectric properties of quartz as distinct from ordinary dielectric properties, and is $\delta V/e$ by application of equation (4). It is obviously in phase with V and cannot take part in the phase variations of the piezo-vibrator in the vicinity of resonance. The ability to vary in phase from the applied P.D. and thus to react on the rest of the circuit is restricted to the remaining part of the strain, namely, $s' - \delta V/e$. This other component of the strain when multiplied by $l'/2$ the distance from node to antinode in the vibrator, gives our x the instantaneous displacement of one end of the vibrator. It should be noted that this is not quite as large as the actual displacement of the end of the vibrating bar for x is to be measured from the non-reacting[11] displacement $l'\delta V/2e$. The latter is the only displacement when the driving

[11] Cady, W. G., Proc. I.R.E., 10, 83, 1922. This is here called "equilibrium elongation."

frequency is remote from resonance or when the plate is highly damped.

The instantaneous velocity of the piezo-vibrator face is $\dot{x}=0.5l'(\dot{s}'-\delta V/e)$ which when solved for $\dot{s}'$ and substituted in (6) gives:

$$I=\dot{V}C'-A\epsilon\delta\dot{V}/e-2\epsilon A\dot{x}/l'=\dot{V}(C'-A\epsilon\delta/e)$$
$$-2\epsilon A\dot{x}/l'=\dot{V}C_1-2\epsilon A\dot{x}/l' \qquad (7)$$

$A\epsilon\delta/e$ thus appears as a small correction to the capacity of the piezo-condenser (about 1 per cent). Considering the two terms of C_1 it is seen by $C'=k'A/4\pi e$ that the effect of $A\epsilon\delta/e$ is to reduce[12] the $k'/4\pi$ by the amount $\delta\epsilon$ which means reducing n' by $\delta\epsilon$, i.e., back to k and n the dielectric constant and suscepti-

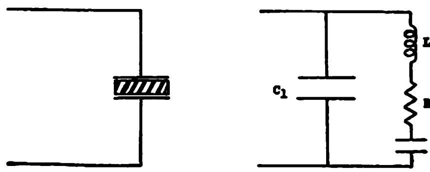

Fig. 1—Piezo-Electric Resonator and Its Equivalent Network

bility of unstrained quartz. We shall then, hereafter, consider the parallel condenser no longer as a piezo-condenser, but as having the capacity of an ordinary dielectric, i.e., as if the quartz dielectric were not piezo-electric, and call it C_1 instead of C'.

Eq. (7) is a statement of Kirchoff's current law, and may be considered either as a vector equation or an equation in instantaneous values. The current to the piezo-vibrator is proportional to and in phase with the rate of change of its modified resonating strain or to the velocity of the vibrating face of the crystal. From the analysis of mechanical resonators of a single degree of freedom[13] the instantaneous velocity $\dot{x}$ of the resonator is expressible in terms of the sinusoidal driving force of maximum value F_0 and the vector mechanical impedance Z of the resonator by Eq. (8) which is of the same form as the current equation in a series electric circuit.

[12] Dye, p. 453 of paper to be later cited, finds experimentally such a reduction in the capacity of the parallel condenser, and points out that his observed decrease is more than here accounted for. The present decrease should enter every a-c. determination of this capacity whether near crystal resonance or not.

[13] Cf. Crandall, "Theory of Vibrating Systems and Sound," p. 9.

$$\dot{x}=\frac{F_0 \cos \omega t}{Z}=\frac{F_0 \cos \omega t}{N+j(\omega M-g/\omega)} \tag{8}$$[14]

Substitution of (8) in (7) gives the Eq. (9) for current to the crystal.

$$I=I_1+I_2=C_1\dot{V}-\frac{2\epsilon A F_0}{l'Z}\cos \omega t \tag{9}$$

When the equivalent driving force on the vibrating face (area A') of the crystal is expressed in terms of the maximum applied field V_0/e by Eq. (4), $F_0=2A'p_0=2A'V_0\epsilon/e$ Eq. (9) takes the form

$$I=I_1+I_2=C_1\dot{V}-\frac{V_0 \cos \omega t}{\frac{el'N}{4\epsilon^2 AA'}+j\left(\frac{Mel'\omega}{4\epsilon^2 AA'}-\frac{el'g}{4\epsilon^2 AA'\omega}\right)} \tag{10}$$

I_2 is the current to the vibrating crystal or to whatever group of circuit elements may be imagined to take its place. The

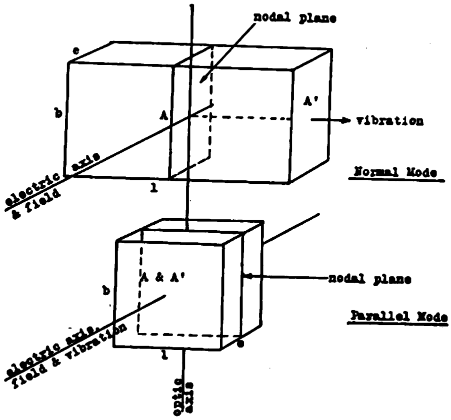

Fig. 2—Crystal Dimensions in Normal and Parallel Fundamental Modes

denominator is, by definition of electrical impedance, (ratio of P.D. to current) the electrical impedance of this imagined branch. This denominator has the form of the impedance of a series chain of electrical elements—inductance L, resistance R, and capacitance C—for which the expression is $R+j(\omega L-1/\omega C)$. Hence the elements are

$$R=\frac{el'N}{4\epsilon^2 AA'};\ L=\frac{el'M}{4\epsilon^2 AA'};\ C=\frac{4\epsilon^2 AA'}{el'g};C_1=\frac{A}{e}\left(\frac{k'}{4\pi}-\epsilon\delta\right)=\frac{Ak}{4\pi e} \tag{11}$$

[14] This is also the derivative of Cady's equation (20). *Phys. Rev.*, 19, 6, 1922.

Thus the entire piezo-electric resonator may be replaced by the equivalent network of Fig. 1 where the condenser C_1 represents the quartz dielectric and the series chain represents the piezo-vibrator.

The values of the elements of the series chain depend upon the dimensions of the crystal, upon its characteristics as a mechanical vibrator, and upon the value of the piezo-electric constant of the quartz. The factor $el'/4\epsilon^2AA'$ is a sort of conversion factor which converts impedance when measured in mechanical units for the vibrator into impedance in electrical units for the current. The R, L, and $1/C$ would all have smaller values and thus have more effect on the current in the external circuit if a piezo-electric material were used which was more strongly piezo-electric, i.e., had a larger piezo-electric constant.

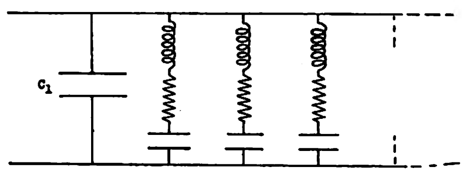

Fig. 3—Single Network Equivalent to Resonator at a Number of Its Natural Frequencies.

The expressions (11) may be simplified for either mode of vibration by substituting for area A' of the moving face, the crystal dimension l' in the direction of the vibration (i.e., the half-wavelength in the crystal) and the area A of the faces which the electrodes cover, their respective values in terms of the length l, breadth b, and thickness e of the crystal. These are measured in directions which are respectively normal to the electric and optic axes, along the optic axis, and along the electric axis. For the *normal mode* the series elements become[15]

$$R=\frac{N}{4\epsilon^2b^2};\ L=\frac{M}{4\epsilon^2b^2};\ C=\frac{4\epsilon^2b^2}{g} \tag{12}$$

[15] The author outlined the derivation of this equivalent network before the American Physical Society in Washington in April, 1925, the abstract appearing in *Phys. Rev.*, 25, 895, 1925. Since then similar equivalent networks have been published. That of Dye (*Proc.* Phys. Soc. of London, 38, 399, 1926) was derived in form from a general dynamical theorem published by Butterworth, loc. cit. and the values of the elements obtained from resonance curves taken for the resonator. When deriving the network the present author was aware of Butterworth's general theorem but preferred to base his derivation on first principles, with fundamental piezo-electric theory as the starting point. The network was first derived from equation (18) of Cady's I.R.E. paper by a few simple steps,

and for the parallel mode

$$R=\frac{e^2N}{4\epsilon^2A^2}=\frac{e^2N}{4\epsilon^2l^2b^2};\quad L=\frac{e^2M}{4\epsilon^2A^2}=\frac{e^2M}{4\epsilon^2l^2b^2};\quad C=\frac{4\epsilon^2A^2}{e^2g}=\frac{4\epsilon^2l^2b^2}{e^2g} \tag{13}$$

The quartz-condenser C_1 has the same value for either mode, for plates of the same area and thickness.

A single network may be drawn (Fig. 3) to represent the crystal in all of its various modes of vibration provided the frequency of each mode is remote enough from other resonance frequencies to insure that the other series chains have very high impedance —a frequency separation of but a tenth of a per cent or so with quartz resonators.

Illustrative values for the elements of the network will now be given. The plate N_2 which was described by Cady[16] and used in his early experiments, is a rod about $30\times4\times1.5$ mm. vibrating at its *normal fundamental* (lengthwise vibration) with a resonant frequency of about 90 kc. per second. Substituting in the formulas (12) and converting from e.s.u. to practical units, the following values are obtained:

$L=137$ henrys	$C=0.0235$ $\mu\mu f$.
$R=$ about 15,000 ohms	$C_1=3.54$ $\mu\mu f$.

For a plate of a size commonly used in oscillators ($25\times25\times2.5$ mm.) vibrating at its *parallel fundamental* (thickness vibration) near to a resonant frequency of about 1100 kc. per second, on substitution in Eqs. (7) the following values are obtained:

$L=0.33$ henrys	$C=0.065$ $\mu\mu f$.
$R=$ about 5,500 ohms	$C_1=1.0$ $\mu\mu f$.

The numerical values for R in the above networks were computed not from the known viscosity characteristics of quartz, but rather from an assumed apparent viscosity ($Q=100$) which is intended to include in a rough way the effects of air damping and mounting friction. If the crystal were mounted in a vacuum and free from mounting friction the viscosity would probably be

—a method which avoids the "modified strain" used in the present development.

Other published papers in which an equivalent network has been shown include:

Balth. van der Pol.—"Gendenboch Uedenlausche Vereeniging voor Radiotelegrafic" (1926, p. 293–8). Abstract appeared in *Jahrb. der. drahtl. tel.*, **28**, 194, 1926.

F. Bedeau, *QST* Francais, **8**, p. 22, 1927.

Y. Watanabe.—*Elektr. Nachr. Techn.* **5**, 45, 1928.

Y. Watanabe.—*Jr.* I.E.E. of Japan, No. 466, 506, 1927.

[16] Proc. I.R.E., loc. cit.

less than half of this value, and the values of R thus only half as large. The value $Q=100$ is that which corresponds to the decrement which Cady obtained for the resonator N_2 vibrating in air with electrodes as close as possible to the crystal.

The assumption of this same value of the viscosity from which to compute R for the longitudinal mode is even less justified because of the entirely different shape of the crystal. Also Cady's theory of longitudinal vibrations of viscous rods is not rigorously applicable to the "thickness" vibrations of a plate, since in the rod the cross-section is supposedly small compared to the length. However, for the fundamental compressional mode the procedure should be valid as a first approximation when the proper elastic modulus is used.

Our application of piezo-electric theory in the above development has involved the assumption that the electrodes were in contact with the quartz. This is usually not the case in the practical application of the piezo-electric resonator. Even when one supposes that the electrodes are in contact, the vibrations of the crystal push them periodically away so that there is always at least a small air gap and perhaps a varying one. The complications which this air gap introduces are treated experimentally by Dye who represents the capacity of the gap by a small capacity in series with a network of the type here derived.

The reactions caused by stationary compressional waves in the air gap should be represented, as Dye[17] points out, by a secondary resonant circuit to be equivalent to their system of vibration as coupled to the piezo-vibrator.

The derived values for the network hold rigorously only in the vicinity of the mode of resonance for which they were derived. This limitation is inherent, as Cady makes clear, in the substitution of the equivalent damped mechanical resonator for the actual viscous rod. But as the series chain has a large positive reactance above resonance, and a large negative reactance below resonance, the network reduces, so far as practical considerations go, to the capacity C_1 except in the vicinity of resonance.

The magnitude of the elements of the network are at first sight rather striking. When the radio engineer considers an inductance coil of 100 henrys or more and tries to picture such a coil having so little distributed capacity as to remain inductive at 90 kc., serious constructional difficulties arise in his mind. To

[1] Loc. cit. p. 428.

tune such a coil to resonance at 90 kc. with a condenser having a capacity of but 1/40 of a $\mu\mu f$. requires that the coil shall have ridiculously small distributed capacity. Fortunately, we are not concerned with the construction of such a coil, for we already have in effect the coil and the condenser tuned to the desired frequency in the properly cut quartz crystal. Furthermore, the coil and condenser whose places the quartz crystal takes are put into so small a space in quartz that we have little worry from stray inductive and capacitive fields. Though these values of L, R, and C seem absurd from the constructional standpoint, nevertheless, to the engineer who wishes values of electrical elements to use on paper in design problems, their magnitude presents no such difficulties.

Marked resonance phenomena seem oddly associated with resistances of the order of 15,000 ohms. It is to be recalled, how-

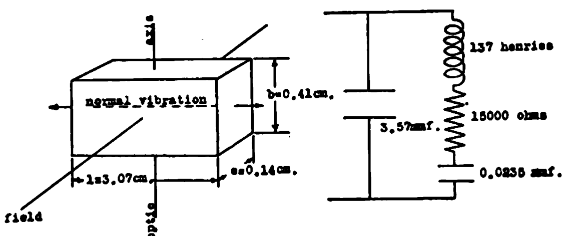

Fig. 4—Resonator N_2 and Its Equivalent Network for Normal Fundamental Mode

ever, that it is the ratio L/R which determines the time constant of a coil and the decrement of a circuit. Here, though R is large, L is even larger in proportion than with most tuned circuits, and the decrement is thus extraordinarily small.

The sharpness of the resonance obtained with the piezo-vibrator may be seen from the ratio of the reactance of L at resonance to R. Consider the plate N_2: 137 henrys at 90 kc. has a reactance of nearly 80 megohms, while the resistance is but about 15,000 ohms; their ratio is about 5,000.[18] At a frequency 1 per cent different from the resonance frequency the reactances of C and L which were equal and opposite at resonance, now

[18] In this connection the "mechanical resonator" described by Moore and Curtis, Bell System Technical *Journal*, 6, 222, 1927, shows a reactance-resistance ratio ten times as large as in the present case. Their resonator was a carefully mounted steel rod. From more recent estimates of the decrement of quartz resonators the reactance-resistance ratio should have a value several times the value of 5000 used.

differ by one hundred times the value of R. Even at 1/10 per cent from resonance their difference is nearly 10 times R. To shift the angle of the impedance of the piezo-vibrator by 45 deg. it it necessary to vary the driving frequency but about 1/100 of one per cent from resonance,—about 9 cycles at 90 kc.

When the frequency is varied about 0.3 per cent above resonance, or less than 300 cycles, the positive reactance of the piezo-resonator is equal in magnitude to the negative reactance of the quartz condenser. At this frequency the mounted resonator has the resistive impedance which is characteristic of parallel resonance (anti-resonance) and is greater than the impedance of either branch. If we call this frequency of parallel resonance f_3 and the frequency of series resonance of the piezo-vibrator f_1, there is still another frequency of resonance in between these.

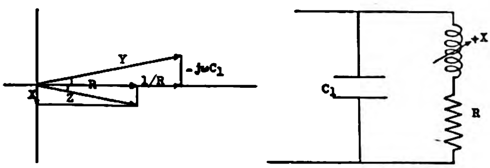

Fig. 5—Impedance and Admittance at f_2

The parallel network, a resistance and a capacity in parallel at f_1, offers there an impedance with a slight negative angle. At a frequency slightly below f_3, however, the positive angled impedance of the L, C, R chain has a lower value than the condenser C_1 by definition of f_3. Hence the parallel network has an impedance whose angle is positive near f_3, and therefore must have changed sign between f_1 and f_3,—say at f_2 a second frequency where the device acts as a pure resistance. Though this frequency may be determined by setting up the equations for the admittance of the parallel network and solving for the condition that its angle shall be zero, it is more easily visualized by vector methods.[19]

The frequency f_2 in question may be shown to be extremely close to f_1, for the network of the crystal N_2 within about 1/30 of one cycle. f_2 may be seen from the vector diagram (Fig. 5) to be that frequency for which the admittance Y for the piezo-

[19] The vector admittance circle as applied by Cady is, of course, best suited to the analysis of the resonator behavior, as it is also for the analysis of the network. Cady's circle is applicable to the network on substitution of circuit elements for crystal elements.

vibrator has such an angle as to result with the admittance $-j\omega C_1$ of the parallel condenser in a combined admittance of zero angle. (For N_2 at 90 kc., $1/\omega C_1$ is about 1/2 a megohm.) The angle of Y is equal, but of opposite sign, to that of the impedance Z of the piezo-vibrator and has a value X/R. (Here X is

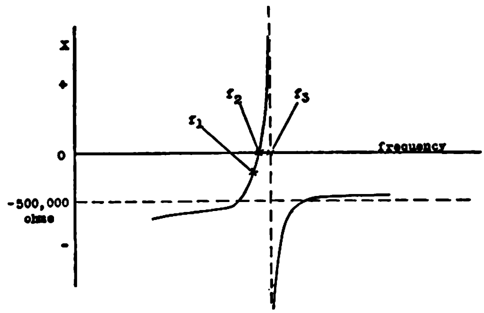

Fig. 6—Reactance of Resonator N_2

the reactance of LCR.) If the angle is small, Z is practically equal in magnitude to R and hence its reciprocal Y to $1/R$. Thus in the present case where R=about 15,000 ohms, $Y=1/15000$ and the angle of Y (or of Z) is (by $\omega C_1/Y$) 0.3 (as $1/\omega c_1=500{,}000$ ohms) from which by equal angles $X/R=0.03$. From the resistance and reactance data given above it may be seen that this ratio of X/R occurs at a frequency only 3/10,000 of one per cent

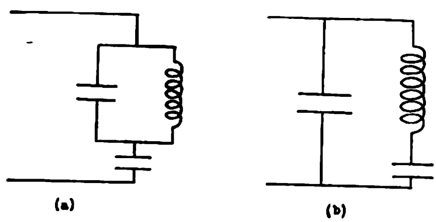

Fig. 7—Two Forms of Network Which Are Externally Equivalent

above resonance,—or 0.3 cycle. The smallness of the angles (1/30 radian) of the impedance and admittance in this illustration justifies the approximations used (angle for sine, and for tangent). The nature of the reactance variations of the piezo-electric resonator as above analyzed for N_2 in the vicinity of resonance is shown qualitatively by the curve of Fig. 6.

Johnson and Shea[20] have demonstrated the rigorous equivalence between networks of the two forms shown in Fig. 7, pro-

[20] Bell System Technical *Jr.*, 4, 60, 1925.

vided the ratio of the resistance to reactance is the same for both coils, and the ratio of resistance to reactance is the same for all condensers. Hence the present network equivalent to the resonator may be converted to the form (*a*) Fig. 7. For the conversion formulas the reader is referred to the paper cited. The equivalence of these two forms is of particular interest in connection with any allowance which should be made for the gap effect as separate from the reaction of the resonant air column. The present network has assumed the electrodes to be in contact with the crystal. Dye[21] shows experimentally that the effect of the gap may be represented by a capacity (which he calls K_2) in series with a network (our Fig. 1). If such a condenser can be placed in series with the network (*b*) Fig. 7 to include the gap, it can also be placed in series with (*a*) Fig. 7 and can then be allowed for by a revision of the value of C_2. From this revised network (*b*) Fig. 7 a new network of type (*a*) Fig. 7 can be drawn, but with values of the constants which are now corrected for the gap. Thus it would seem to be unnecessary in picturing the circuit behavior of the resonator to introduce a series condenser for the gap effect. The air column effect, however, as was pointed out above must be considered to involve a separate tuned circuit coupled in some way to the present series *LCR* branch.

The author has used the cathode-ray oscillograph to demonstrate experimentally the nature of the variations in the crystal impedance as the frequency is varied through resonance. Using as the oscillograph the Western Electric[22] vacuum-tube No. 224–A, one pair of deflecting plates of the oscillograph is arranged to show the P. D. across the resonator while the other pair of plates shows the current to the resonator. Since the oscillograph has a high impedance, resistances are placed across each deflecting system. Thus the deflections and their phases are not characteristic of the mounted crystal alone but depend also upon the magnitude of the resistances or condensers used. Bridge arrangements of the types shown as (*a*) and (*b*) (Fig. 8) have been used. Fig. 9 shows a typical set of ellipses sketched for the series of frequencies shown in the vicinity of resonance (normal mode) about 90 kc. for a plate ($30 \times 10 \times 1$ mm.). The elements of the bridge were all resistances; those across the deflecting plates being

[21] Dye, loc. cit.
[22] The circuit used is analogous to that described for studying dielectric losses in the "Instructions for Operating the No. 224-A Vacuum-Tube."

25,000 ohms each, and the third element 100,000 ohms. As the crystal frequency is approached from the low-frequency side, the ellipse shifts through the phases shown, all in a narrow frequency range in the vicinity of resonance. For the figures shown the deflections up and down the paper are proportional to currents through the crystal, and those across the page to P.D. across crystal and resistance in parallel.[23]

One difficulty in following such phase changes continuously through resonance in early experiments was the reaction of the crystal back on the source,—its stabilizing action. The sudden changes in phase of the load react in such a way on the source as

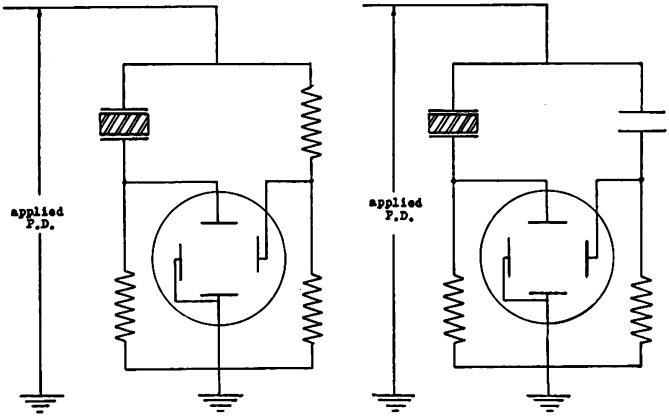

Fig. 8—Piezo-Electric Resonator in Oscillograph Bridge

to prevent continuous tuning through the resonance frequency of the load unless the crystal forms a very insignificant part of that load. In the present experiments the driving oscillator is of such power and so loosely coupled to the crystal and oscillograph bridge that no reactions of this sort are detected. As a result the ellipses vary continuously through the sorts of changes indicated as the frequency is slowly varied through resonance.

A series of figures of ellipses of this type has been made for various sorts of quartz resonators varying in frequency from 20 kc. to 1000 kc. It is hoped that careful measurements on these ellipses taken at carefully determined frequencies in the vicinity of resonance can be solved to yield reliable estimates of the magnitude of the elements which make up the equivalent network of

[23] Watanabe, *Jr.* I.E.E. Japan, May 1927, shows similar oscillograph figures for the piezo-resonator.

the resonator, though no estimate of the precision to be expected has yet been made.

The use of the cathode-ray oscillograph is found of further value in distinguishing between true and false response frequencies of the crystal. When a crystal is placed across a vacuum-tube oscillator and the oscillator is tuned continuously through a range of frequencies, crystal reactions result in clicks heard in phones connected to the oscillator or in a secondary circuit. Such clicks are heard not only for oscillator frequencies which match crystal vibration frequencies, but occur as well for those oscillator frequencies which have harmonics of appreciable magnitude at a crystal frequency. The crystal is thus driven by

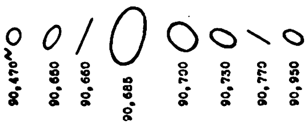

Fig. 9—Oscillograph Patterns Near Resonance

the harmonic, and a click resulting from its reaction on the oscillator circuit may be heard. These clicks are usually not so loud in this latter case as when caused by the fundamental frequency of the oscillator, and a basis for their recognition is had in this difference in loudness. On the other hand there are marked differences in the loudness of clicks for different true crystal resonance frequencies. With the cathode-ray oscillograph, however, when the crystal is responding to the driving oscillator, the ellipses shown above appear if the response is to the fundamental of the oscillator, but if to some over-tone in the oscillator output, the patterns show a two-ring, or a three-ring structure instead of the simple ellipse. This method has been used to locate the resonance vibration frequencies of resonators of various shapes. It has the added advantage of permitting one to explore carefully the region in the immediate vicinity of a resonance frequency. Thus it has been found in some cases that a crystal has a pair of resonance frequencies so close together that by the click method they have appeared as one.

The author cannot close this paper without acknowledging Professor Cady's patient counsel and helpful assistance in its preparation.

SOME CORRELATIONS OF RADIO RECEPTION WITH ATMOSPHERIC TEMPERATURE AND PRESSURE*

By

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Summary—*Night reception and temperature at the receiver are found directly related, maximum reception being associated with maximum temperatures and vice versa. This is the reverse of the relation previously found by Austin for day reception, where falling temperature improved reception, and is therefore another case of the already established inverse relation of night to day reception. The temperature effect appears to be local to the receiver, for no definite relation was found between temperature at the transmitter and reception. A correlation between night reception and pressure was also found, signal strength increasing as areas of low pressure passed over the receiver, and decreasing with the passages of high pressures.*

THE first definite correlation between radio reception and any measured atmospheric state was made by Austin in 1924[1]. Comparing air temperatures with day reception at Washington from low-frequency stations in New Jersey he found an inverse relation, which was particularly marked with cold waves. Early in my measurements of night reception in the broadcast band I examined pressure gradients between transmitter and receiver without finding a definite relation to reception.[2] But when a homogeneous series of reception measurements extending over a period of two years had accumulated, meteorological relations began to appear, particularly those with temperature and pressure which form the subject matter of this paper.

Two series of reception measurements were available for this comparison; my own of station WBBM at Chicago as received in Newton Centre, Massachusetts, and those of Mr. Howell C. Brown at Pasadena, California, of transmission from station KWFI at San Francisco. Mr. Brown's measurements are of peculiar value in this work, as his transmission path (perhaps because it has a considerable component parallel to the Earth's magnetic field) appears but little affected by solar activity and magnetic disturbances. Unfortunately, although my measure-

* Original Manuscript Received by the Institute, April 24, 1928.

* Presented before the International Union of Scientific Radiotelegraphy, Washington, D. C., April 19, 1928.

[1] Proc. I.R.E., 12, 681, December, 1924; 14, 781, December, 1926.

[2] Proc. I.R.E., 15, 95, February, 1927.

ments over a West-East path now cover a period of over two years, the California series began in July, 1927, and is as yet too short for a really satisfactory comparison.

In order to eliminate the effect of seasonal and other long period changes in reception the individual night fields were divided by their moving 27-day average, thus giving a series of index numbers representing the percentage deviation of each night from the 27-day average centering on that night. Selecting a sufficient number—about thirty—of temperature or pressure changes as epoch dates, a mean of the reception index numbers was made for the day of the change and also for each of a number of days before and after.

Daily mean temperatures taken by the Weather Bureau offices nearest the transmitter and receiver are used throughout.

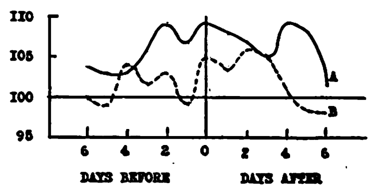

Fig. 1—Newton Reception from Chicago Against Chicago Temperature Changes.

For the Chicago-Newton path, the temperatures are those of Chicago and Boston, while for the San Francisco-Pasadena transmission, those of San Francisco and Los Angeles are employed. To eliminate the effect of small and purely local temperature changes, only changes of 10 deg. F. or over were taken for the Chicago-Newton transmission, and because of the disturbing effect of the East wind on Boston summer temperatures only the fall, winter, and spring months were used. For the San Francisco-Pasadena path, the more uniform California climate made it necessary to take smaller temperature changes, those of 5 deg. F or over being used.

In Fig. 1 is shown a comparison of Newton reception with temperature changes at the Chicago transmitter over a period of thirteen days centering on the days of the temperature change, the ordinates being reception percentages. The full-line curve *A* is of reception accompanying falling temperatures, while the dotted curve *B* is for rising temperatures. No definite relation appears, and the amplitude is small, about five per cent in each case.

But when daily mean temperatures at Boston, 11 kilometers from the receiver, are taken, a more definite relation is found. In Fig. 2 the full-line curve C is for cold waves, while the dotted

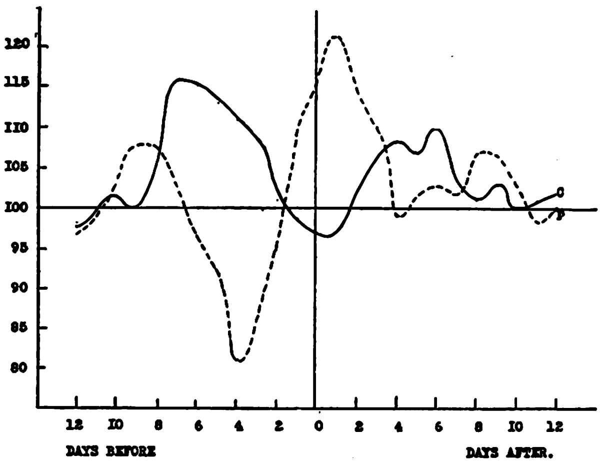

Fig. 2—Newton Reception from Chicago Against Boston Temperature Changes.

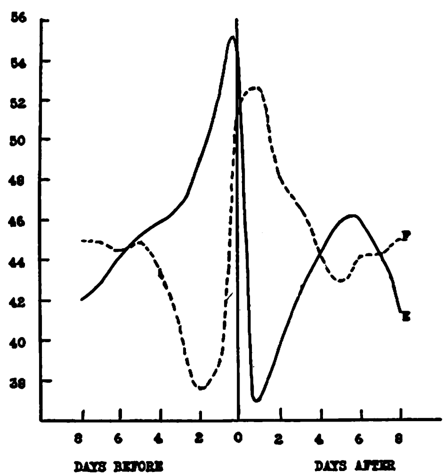

Fig. 3—Character of Boston Temperature Changes.

curve D is for abrupt temperature rise. Considering for the moment only the central part of the figure it will be seen that whereas reception shows a distinct minimum centering on the day of the cold wave, it also shows a marked maximum nearly coinciding with the day of the temperature rise.

To interpret Fig. 2 fully it is first necessary to determine the character of the temperature changes for several days before and after the abrupt rise or fall. In Fig. 3 is shown the result of

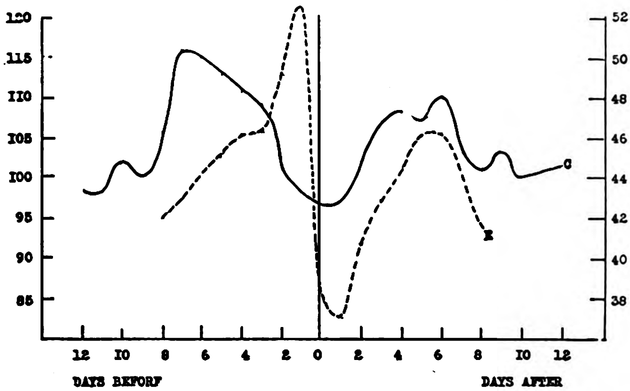

Fig. 4—Comparison of Newton Reception and Boston Temperature. Changes.

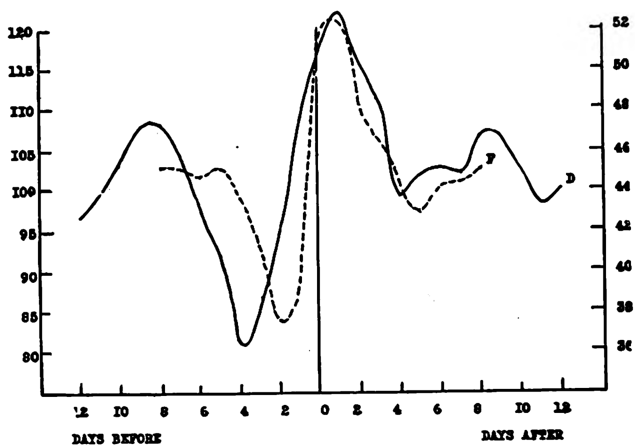

Fig. 5—Comparison of Newton Reception and Boston Temperatures. Temperature Rise.

averaging daily mean temperatures over a period of seventeen days centering on the day of the change, full-line curve *E* being for cold waves, while dotted curve *F* is for sudden rises. The curves are very nearly inverse, showing in each case a slow rise or fall, a sudden fall or rise on the day of the change, and then a slow recovery to normal.

We will first compare reception and temperature before, during, and after a cold wave. In Fig. 4 this is done by superposing reception and temperature curves *C* and *E*, the ordinate scale at

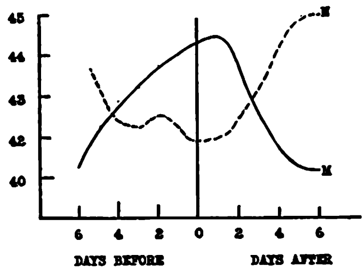

Fig. 6—Boston Temperatures Accompanying Newton Reception Maxima and Minima.

the left being reception percentage, while that at the right is temperature. Clearly temperature and reception are here directly related, although the curves are somewhat displaced, reception changes coming in advance of temperature.

An even clearer relation is shown for temperature rise and reception in Fig. 5. Here curve *D* is for reception, while curve *F*

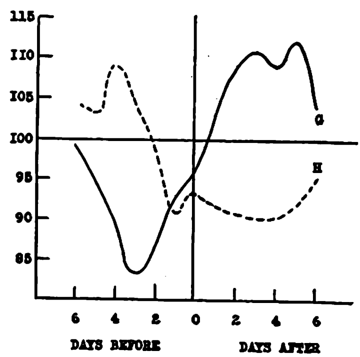

Fig. 7—Newton Reception Against Passage of Cyclones and Anticyclones over Massachusetts.

is for temperature. And in the left-hand portion of this figure the same tendency of reception to lead temperature is shown, which suggests that reception may not be directly related to temperature, but perhaps instead to a temperature-controlling cause.

As a check upon the foregoing, I have reversed the process so far employed, by taking the maxima and minima of my night

reception series as epoch dates, and finding the associated Boston temperatures. The result is shown in Fig. 6, where full-line curve *M* represents the temperatures centering on reception maxima, and dotted curve *N* those accompanying reception minima. Again maximum temperatures are found associated with maximum reception, with minimum temperatures accompanying minimum reception.

There is another meteorological event which repeats at sufficiently frequent intervals for correlation with my night reception data; the passage over Massachusetts of the centers of cyclones and anticyclones, which are the familiar "Lows" and

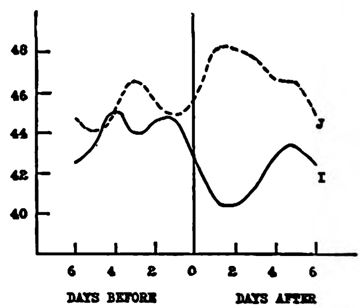

Fig. 8—Boston Temperature Changes Accompanying the Passage of Cyclones and Anticyclones.

"Highs" of the weather map. In Fig. 7 is shown in full-line curve *G* the relation between reception and the passage of Lows, while curve *H* is the relation to anticyclones or Highs. A very distinct inverse relation exists between these two curves, reception showing a decrease before and a rise after the passage of an area of low pressure, while reception is better before and worse after the passage of an area of high pressure.

But cyclones and anticylcones are related to temperature changes, which makes it possible that the relation shown in Fig. 7 is with temperature rather than with pressure. In Fig. 8 is shown in full-line curve *I* the temperature changes at Boston accompanying Lows, while dotted curve *J* represents the temperatures associated with the passage of Highs. A comparison of this figure with the preceding one makes it evident that the temperatures accompanying cyclones and anticyclones are not responsible for the reception changes. The passage of a Low, for example, is associated with a temperature drop a day or two later,

whereas reception, which is directly related to temperature, shows a maximum several days after the passage. Similarly, the passage of an area of high pressure is associated with a temperature rise reaching a maximum a day or two after, but reception is at a minimum during and after the passage.

Although the San Francisco-Pasadena series is as yet too short for satisfactory correlation with meteorological elements, I have compared San Francisco temperatures with Pasadena reception and have found no relation, whereas temperatures taken at Los Angeles, only 18 kilometers from Pasadena, show a distinct direct relation to reception. In Fig. 9 full-line curve *K* is for

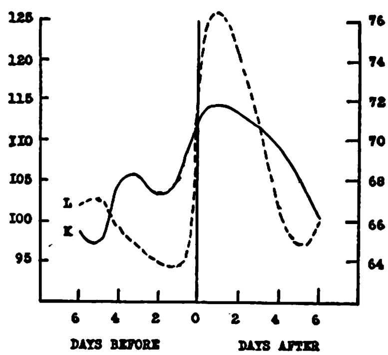

Fig. 9—Pasadena Reception from San Francisco Against Los Angeles Temperature Changes. Rise in Temperature.

Pasadena reception, while dotted curve *L* is for the accompanying temperatures; a clear direct relation is shown.

Through the kindness of Mr. William K. Aughenbaugh, of Altoona, Pennsylvania, I have received a nightly estimate of signal strength in the 3.75 megacycle amateur band since the middle of October, 1927. Although this series is still far too short for comparison purposes, I find in it a distinct direct relation between reception and Altoona temperatures.

These correlations form an interesting supplement to Austin's original discovery, in that they are the reverse of his results. Just as in the past we have found that solar activity and magnetic storms are associated with improved day and depressed night reception, it now appears that cold waves are related to improved day and lowered night reception; it is another instance of the inverse relation between day and night transmission.

It must be understood that these relations between night reception, temperature and pressure do not hold over long periods, as for example between monthly means of reception and temperature. At least in years of sunspot minima night reception is dictinctly better in winter than summer, but this is not a temperature effect; rather it is because of the longer hours of sunlight and increased absorption by vegetation during the summer months.

Inasmuch as both temperature and pressure are related to solar activity[3] it is not safe to assume that the correlations given above are purely those of cause and effect. And it is possible that changes in the troposphere are associated with events above the isothermal layer, either directly or as common effects of a solar cause. Certainly it is difficult to see how changes of temperature, pressure or humidity can in themselves affect radio transmission.

[3] "Solar Radiation and Weather or Forecasting Weather from Observations of the Sun", H. H. Clayton, Smithsonian Miscellaneous Collection, 77, No. 6, June 20, 1925; "Solar Activity and Long Period Weather Changes", H. H. Clayton, Smithsonian Miscellaneous Collection, 78, No. 4, September 30, 1926.

TECHNICAL CONSIDERATIONS INVOLVED IN THE ALLOCATION OF SHORT WAVES; FREQUENCIES BETWEEN 1.5 AND 30 MEGACYCLES *

By

LLOYD ESPENSCHIED

(American Telephone and Telegraph Company, New York City)

THE information given on the accompanying chart is based on the experience of engineers of the Bell System, combined with information secured by engineers of other organizations, and represents what is believed to be the consensus of present knowledge on this subject. The chart was ori-

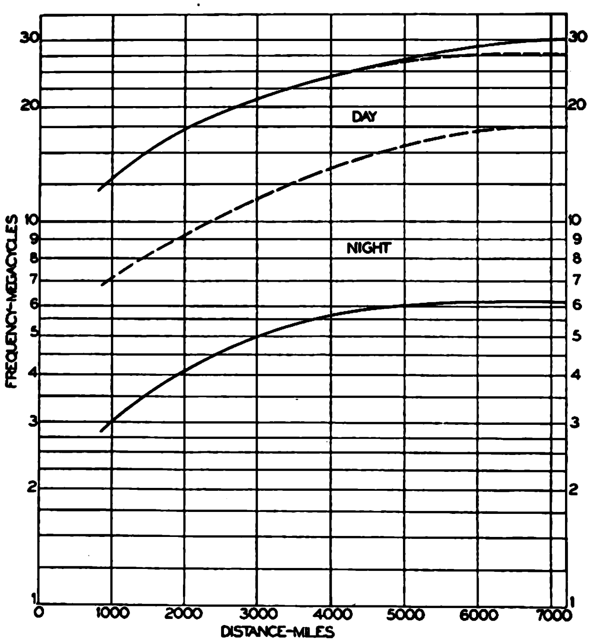

Fig. 1—Approximate Relation of Optimum Frequency to Distance in Short Wave Radio Transmission. 1½ to 30 megacycles, 0 to 7000 miles.

ginally drawn up as something which would be of interest to the Federal Radio Commission and was submitted to the Commission at the hearing upon the allocation of short waves which was held in Washington, January 17 and 18, 1928.

* Original Manuscript Received by the Institute, March 28, 1928.

find there exists the general relationship between frequency and distance delineated by the curves in Fig. 1.

These curves illustrate phenomena well-known to those familiar with short-wave transmission,—the fact that frequencies which may be suitable for transmission to a given distance at night may be quite unsuitable for daytime communication over the same distance, and vice versa; and that, in general, there are definite limitations in the distance for which the various parts of the short-wave radio spectrum are best adapted at a given time. The curves on this chart should be regarded as general boundaries outside of which it is not usually advisable to choose frequencies. For example, to communicate over a distance of 4000 miles during the day time, it is desirable to select a frequency between about 13.5 and 24 megacycles, preferably one near the middle of this range.

National and International Aspects

It is convenient to divide the short-wave spectrum into three bands as indicated at the extreme right of Fig. 2. Of course, these bands are not discrete but merge into one another.

Band A—1500 to 6000 kc. (200 to 50 meters). This band is best adapted to communication over moderate distances in the world-wide sense, distances up to, perhaps, 1000 miles at night. The band may, therefore, be considered as regional in its service range. The higher frequencies may cause interference over intercontinental distances at night.

Band B—6000 to 15000 kc. (50 to 20 meters). In its range for communication this band is more or less regional for that portion of the globe which is in daylight, but may include practically the entire hemisphere which is in darkness. During the winter months of the year the daylight hours for the northern hemisphere of the globe are relatively short and the higher of these frequencies are very widespread in their effect during the winter season.

Band C—15000 to 30000 kc. (20 to 10 meters). This band (the higher limiting frequency being somewhat uncertain) appears to be world-wide in its communication range, extreme distances being reached especially over the hemisphere which is in daylight.

For all three bands world-wide coordination is, of course, necessary in respect to the *services* for which the frequencies are used. Futhermore, world-wide coordination in the individual channel assignments to stations will be required for Bands *B* and *C*, and probably also for the higher frequencey

end of Band *A*. For the lower frequency end of Band *A* it may prove to be practical in large isolated areas, as in the North American Continent, to make individual station assignments without requiring coordination with similar assignments in other continents.

International Frequency Allocation

In the center of the chart are two columns which give the allocation of frequency bands in this range adopted by the International Radiotelegraph Conference, held in Washington in 1927. This allocation becomes effective, for the countries which ratify the Washington Convention, on January 1, 1929. It is, therefore, to be expected that the various national agencies will immediately begin to use this allocation as a guide in making frequency assignments to stations in this range.

Available Frequency Channels

The number of channels which can be simultaneously occupied by radio stations is definitely limited. In the chart are given figures showing the limiting number of channels for radiotelegraph and radiotelephone communication, assuming channel spacings of 1,000 cycles and 10,000 cycles, respectively. This number is far in excess of the number which can be used practically in the present art.

An estimate based upon present general practice is also given on the chart. Such an estimate will vary considerably, depending upon the engineering assumptions. The present estimate shows something less than 1000 channels, either telegraph or telephone, the limitation being not in the band widths of the channels themselves, but in the separation at present required between them to avoid interference. For services which undertake to be continuous and reliable in operation it must be recognized that two or three or four frequencies per station will be required to cover the various parts of the day and seasons of the year.

The more important of the factors which necessitates a substantial separation between channels are outlined below. The separation may be expected to be diminished in time, as the results of further development work are embodied in practical operation.

(1)—Variation in the frequency of the transmitting station. Current practice at the best stations is to use piezo-electric crystal control with temperature regulation. Without such control and regulation, wide variation in the transmitted frequency results.

(2)—Lack of selectivity of receiving equipment. Any radical improvement in the selectivity of short wave radio receiving equipment is likely to involve complicated apparatus and will, therefore, be somewhat expensive. Development along this line must, however, be expected if the most adequate use is to be made of space in this range.

(3)—Practical factors concerned with the use of these channels, such as the geographical relation between sending and receiving stations, the different types of services required and the divers practices of the operating organizations involved.

THE NAVY'S PRIMARY FREQUENCY STANDARD*

By

Robert H. Worrall and Raymond B. Owens

(Naval Research Laboratory, Washington, D. C.)

Summary—In this paper is described a constant frequency source which is used as the Navy's primary working standard of frequency. This frequency source is a crystal-controlled oscillator of special design which was developed at the U. S. Naval Research Laboratory, Washington, D. C. This oscillator employs a type of circuit which is rich in harmonics, over 200 of which may be used as fixed standard frequencies.

It also describes in detail the method employed to determine the fundamental frequency of this standard in terms of Naval Observatory time to an accuracy of about one part in 100,000. The ultimate standard of frequency is therefore the mean solar day.

History

SEVERAL years ago when the Navy began to experiment with high-frequency continuous-wave radio communication, the Navy standard of frequency was an absorption type of wavemeter. It consisted of a precision condenser and a number of coils designed to have a constant inductance and a low radio-frequency resistance. A resonance indicator, if desired, could be loosely coupled to this circuit. This standard wavemeter was calibrated by means of the multi-vibrator or other oscillator the output of which was rich in harmonics. The multi-vibrator in turn was adjusted to resonance with a standard 1000-cycle tuning fork. By this means the wavemeter calibration could be maintained by frequent checking to an accuracy of one-tenth of one per cent. For spark and low-frequency continuous-wave transmission, precision of this order was satisfactory.

One of the first results of the advent of high-frequency transmission was to indicate the inadequacy of a frequency standard the calibration of which might be in error by 0.1 per cent in either direction. For at a frequency of 10,000 kilocycles, this

* Original Manuscript Received by the Institute, January 9, 1928; Revised Manuscript Received by the Institute, March 30, 1928.

* Read by Dr. L. P. Wheeler for the authors at meeting of the International Union of Scientific Radiotelegraphy in Washington, D. C., October 13, 1927.

would mean a possible error of 10 kilocycles, plus or minus. Thus at this frequency a transmitter would occupy a band 20 kilocycles wide, which would increase interference between stations, or greatly decrease the number of communication channels by requiring a greater separation in kilocycles between assigned frequencies. Naturally a service wavemeter calibrated from this standard might conceivably be in error by a much larger per cent than the standard itself.

Therefore there was presented to the Naval Research Laboratory for solution a three-fold problem: (1) to develop for the Naval service a frequency standard whose constancy should be as nearly absolute as possible; (2) to provide a means of obtaining direct from this standard a large number of fixed frequencies for calibration purposes; (3) to develop a method for determining the frequency of this standard in terms of observatory time to better than 0.01 per cent, and as near 0.001 per cent as possible.

Crystal-Controlled Oscillators

The application of the piezo-electric properties of quartz crystals to control the frequency of vacuum-tube oscillating circuits, regardless of small changes in capacity or inductance, suggested a crystal-controlled oscillator as the solution to our first problem. After experimenting with different types of circuits and making a number of refinements in crystal holders a crystal-controlled oscillator was developed which with automatic temperature control was found to give an output of practically absolute constancy.

Tests have shown that the combined effect on the frequency of the standard oscillator caused by five per cent change in plate voltage is less than 0.001 per cent. A ten per cent change in filament voltage produces no noticeable change in frequency. Changing from one tube to another of the same type produces a change in frequency varying from zero to a maximum of about 0.002 per cent. On account of the lack of uniformity in tube constants, several tubes that give the same frequency as the oscillator tube used in the calibration are kept as spares, which practically eliminates the error due to tubes. The temperature coefficient of the standard crystal in its holder is less than one part in one hundred thousand per degree centigrade.

Types of Circuits

One type of circuit used in the preliminary work was that shown in Fig. 1. This tuned plate circuit,[1] using a small inductance and a relatively large capacity, gives maximum output on the fundamental frequency and a rapidly diminishing amount of energy on the harmonics. In other words, this is a transmitter type of circuit, and only a few harmonics could be

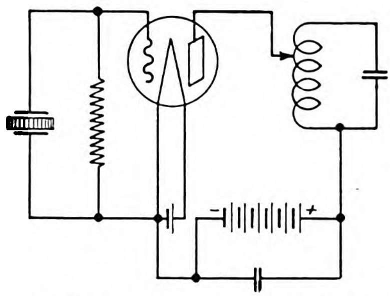

Fig. 1—Tuned Plate Type of Crystal Oscillator Circuit, Having a Low *L/C* Ratio.

used for calibration purposes. This circuit was therefore modified with the purpose of reducing the strength of the fundamental and intensifying the harmonics, so that each multiple of the crystal frequency could be used as a standard frequency. This

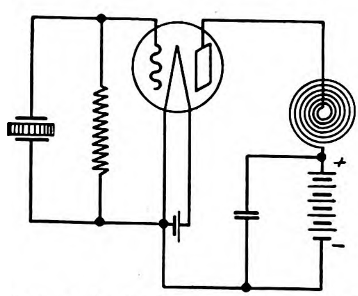

Fig. 2—Circuit Used in Navy Standard Oscillator to Develop Harmonics of Crystal Frequency

modification consisted in replacing the tuned plate circuit, with a very large inductance having a small distributed capacity and no condenser across the coil. See Fig. 2. This untuned coil must have a natural frequency higher than the crystal fundamental; that is, it must be an inductive reactance at the crystal

[1] Developed by Dr. J. M. Miller.

frequency to give the feedback in the right phase relation to the grid potential to sustain oscillation.

To bring out more clearly the difference in the strength of harmonics in these two types of plate circuits, it may be stated that the amount of radio-frequency current in the plate circuit of an oscillator on a given harmonic frequency depends greatly on the ratio of the external impedance at this frequncy to the internal plate impedance of the tube. When these impedances are about equal for a given frequency, we get the maximum energy on this frequency. If the external circuit impedance is small compared to the tube impedance at a given frequency the energy radiated at this frequency will be correspondingly small.

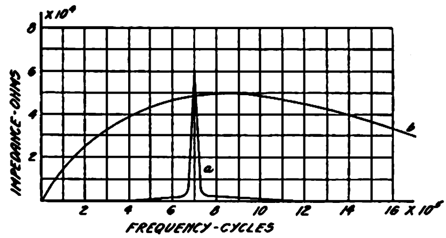

Fig. 3—Impedance Curves (calculated) for Circuits in Figs. 1 and 2.

Now the tuned circuit with a low L/C ratio gives an external impedance of about the order of the tube plate impedance, or greater, over a narrow band of frequencies. See Fig. 3 (a). For instance, the external impedance is relatively very small at a frequency five times the crystal fundamental. Therefore this harmonic would be very weak in comparison with the energy radiated on the fundamental frequency.

On the other hand, a coil with a large L/C ratio gives an impedance curve that is very broad, indicating that the strength of the fundamental and of a large number of successive harmonics will be of the same general order, as indicated in Fig. 3 (b).

This oscillator may be considered as a very low-power crystal-controlled transmitter radiating on perhaps two hundred frequencies each one of which is an integral multiple of the crystal fundamental frequency.

To make use of all these harmonics, as in the calibration of heterodyne frequency meters, it is necessary to couple this oscillator, as well as any oscillator to be calibrated from it, to a detector and amplifier. The crystal oscillator and the detector

and amplifier are built into a single unit, which with the automatic temperature control device is called the *crystal-controlled standard oscillator*. See Fig. 4. The input of the detector is untuned and the coil is designed to respond to all frequencies with-

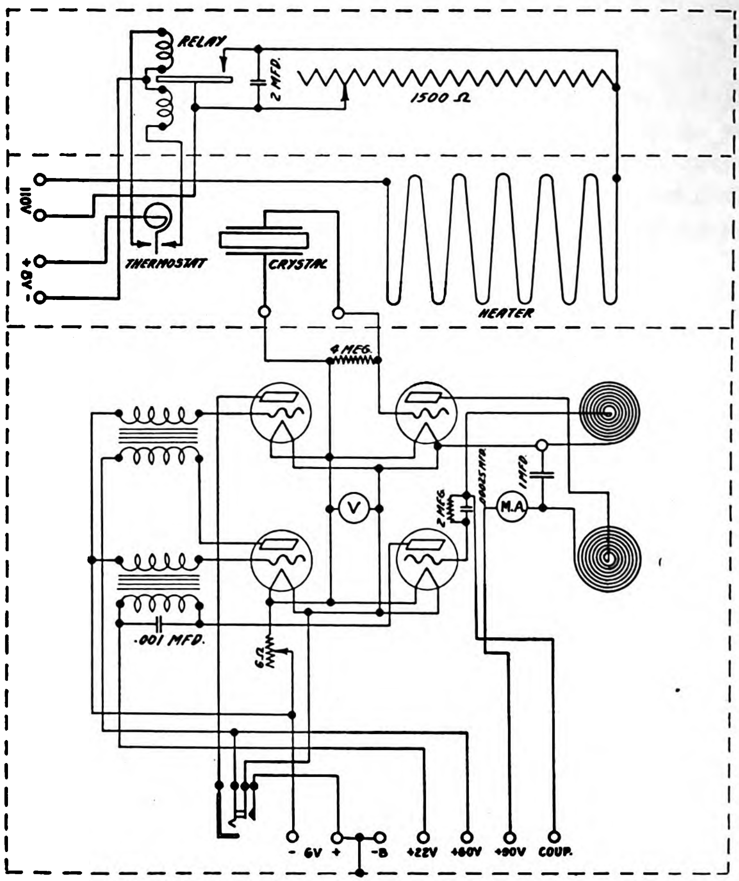

Fig. 4—Circuit of Crystal-Controlled Standard Oscillator Showing Heater Circuit.

out any adjustments. In practice, the coupling is fixed between the crystal oscillator circuit and the detector coil, and a coupling wire is run from the detector sufficiently close to the oscillator under calibration for coupling purposes. When the heterodyne frequency meter (or other type of continuously variable oscillator) is adjusted to within an audible frequency of any har-

monic of the cystal oscillator, the beat note between these is rectified by the detector and amplified by one or two stages of amplification.

The decrease in beat note strength between the successive harmonics of the crystal fundamental frequency and another oscillator is too slight to notice. The signal produced by the hundredth harmonic is of the same general order of audibility as that caused by the fiftieth, and the two hundredth harmonic gives a note of easily readable strength. This device is portable as illustrated in Figs. 5 and 6. At the Naval Research Laboratory the frequency of the standard crystal is 25 kc. From this one crystal and the associated circuits as just described, standard

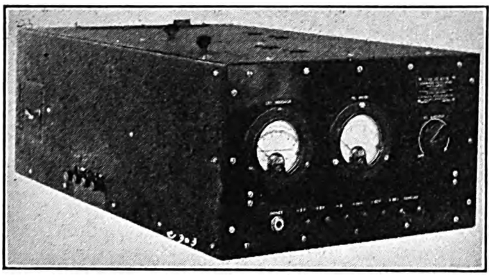

Fig. 5—Outside View of Navy Standard Oscillator.

frequencies are obtained every 25 kc. up to 5000 kc. For a high-frequency range, another standard is provided with a crystal of a frequency of 250 kc. These two frequency standards enable us to cover the entire radio-frequency scale.

Calibrating the Standard Oscillator

General Description of the Method Used.

Now that a constant frequency source had been developed, capable of giving approximately two hundred points for calibration purposes, the final problem was to provide a means of determining the frequency of this standard to as near 0.001 per cent as possible.

The method that was finally decided upon lends itself particularly well to the calibration of low-frequency sources, such as the 25 kc. standard, but it is not limited to any such frequency. In brief, the method employed makes use of a motor-driven generator the frequency of which is varied until some harmonic of the generator output gives "zero beat" with the fundamental of the

crystal oscillator. Then the generator is operated at this exact frequency for some minutes during which time a record consisting of a dot per second is imprinted upon a sheet of paper on a revolving chronograph drum through the medium of a striker actuated by an electrical circuit operated by a standard chronometer. From the slope of the lines of dots thus made on the chronograph drum, the frequency of the generator is very accurately determined. This frequency, multiplied by the number of the harmonic of the generator which was adjusted to the

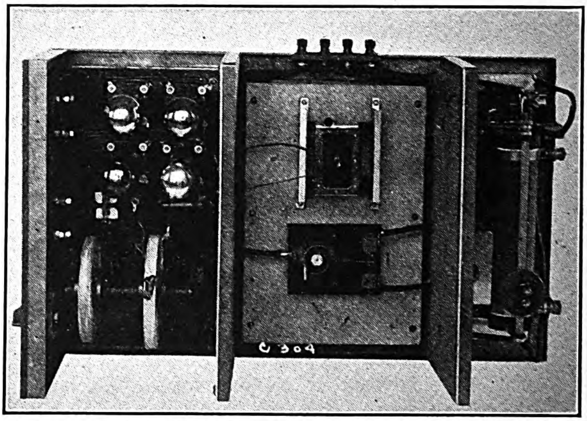

Fig. 6—Standard Oscillator with Cover Removed.

"zero beat" with the crystal oscillator, gives the frequency of the oscillator itself.

To make clear how the method works out in practice, a more detailed description of the apparatus and procedure follows:

Apparatus Used in the Calibration of the Crystal-Controlled Standard Oscillator.

(1) A 500-cycle motor generator.

(2) A very accurate chronometer.

(3) A chronograph.

(4) A Maxwell bridge curcuit.

(5) A circuit for the suppression of the 500-cycle fundamental of the generator and the accentuation of the harmonics of the 500-cycle fundamental.

(6) A special amplifier.

(1) The motor generator outfit consists of a five horse power d-c. motor connected by a heavy fly wheel to an alternator

type 500-cycle generator, operated on a high capacity 110-volt battery. (See Fig. 7).

(2) The chronometer is a large Navy standard chronometer adjusted and periodically inspected at the Naval Observatory. The chronometer is checked daily at the Laboratory against the standard time signals. The original rate as determined at the observatory was one and one-half seconds a day. Our observations covering several years show a slight seasonal variation in the rate, about one second a day in summer and one and one-half

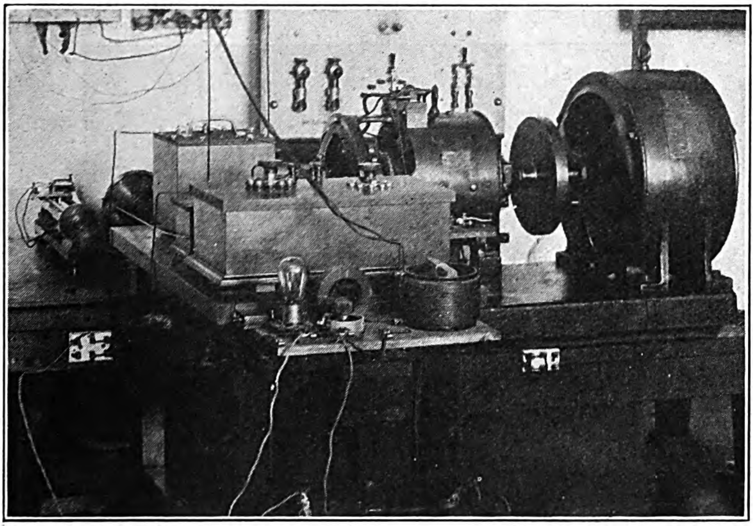

Fig. 7—Apparatus Used in Calibrating Crystal. Motor-Generator at Right, Maxwell Bridge Units in Center, Chronograph Drum at Left Rear.

seconds in the winter. A correction chart has been plotted showing the rate for any time of the year. But when calibrations of the standard crystal are being made, the rate is determined from the daily check against the time signals.

(3) The chronograph consists of the following units; the chronometer, a relay, a striker and a drum. The drum, Fig. 8, is a hollow brass cylinder about 7 in. in diameter and 16 in. long, accurately machined and geared to the generator by means of reduction gears, giving one rotation of the drum to 150 revolutions of the generator shaft. On this cylinder is placed a piece of paper upon which the striker referred to above makes a record as it moves along the face of the drum through the medium

of a worm. The striker records a dot upon the paper once a second (omitting the 59th second every minute) and is actuated by the chronometer. One rotation of the generator shaft corresponds to an alternating potential output of twenty cycles, as can be noted by counting the poles. Now when the drum, operated through the reduction gear just mentioned, rotates once in exactly six seconds, as indicated by six rows of dots exactly parallel to the axis of the drum on the record sheet, the generator frequency is $\frac{150 \times 20}{6}$ or 500 cycles per second.

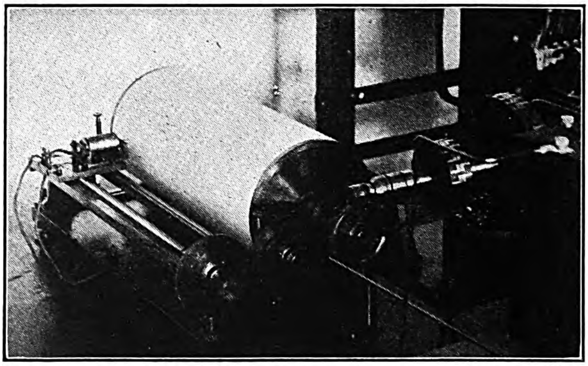

Fig. 8—Chronograph Drum, with Paper Mounted for Calibration Record. Recording Relay at Left, on Carriage.

If the rows of dots slope up the generator output is less than 500 cycles; if they slope down, it is more. From this chronograph record the exact generator frequency during a run is determined, as will be explained in detail further on.

(4) The Maxwell bridge, a diagram of which is shown in Fig. 9, is used in connection with the calibration of the standard crystal as an aid in maintaining the speed of the motor-generator constant at the value desired.

The condenser shown in one arm is charged and discharged from the battery across the bridge by means of a rotating interrupter driven from the generator shaft. The commutator segments of the interrupter operate through three brushes to charge and discharge the condenser.

When the bridge is balanced at a given generator speed, the slightest variation from this speed causes a deflection of the beam of light reflected from the galvanometer mirror on a ground

glass scale. The direction of the movement of the light beam indicates whether the generator is running too slow or too fast, thus enabling the operator to vary the pressure on the fly wheel to restore the generator speed to the zero beat condition.

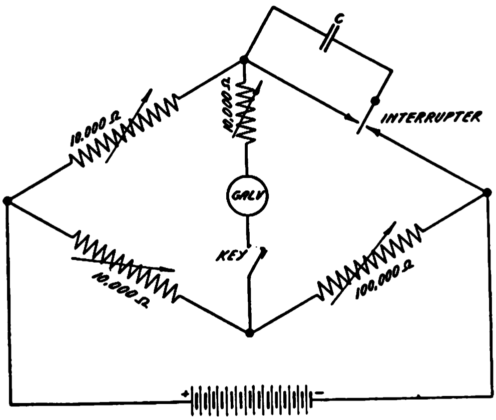

Fig. 9—Maxwell Bridge Circuit.

This use of the generator, chronograph, and Maxwell bridge was patterned after the apparatus used at the Bureau of Standards[2] for the accurate measurements of small capacities.

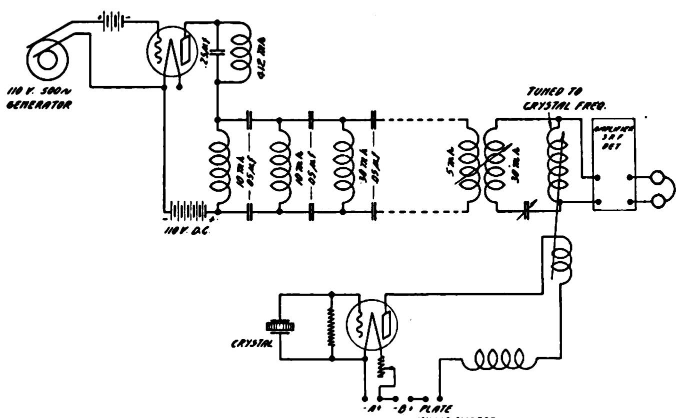

Fig. 10—Crystal Calibrating Circuit.

(5) The method of obtaining harmonics from the output of approximately 500 cycles is to impress this frequency upon the grid circuit of a vacuum tube, the plate circuit of which is

[2] Bulletin of the Bureau of Standards, vol. 3, No. 4.

rich in the required harmonics. As can be seen from Fig. 10 a tuned resonant circuit in the plate circuit partially absorbs the 500-cycle fundamental which, of course, is very strong. The rest of the circuit consists of a filter designed to pass only the high-frequency harmonics of the generator fundamental. This filter circuit terminates in a coupling coil which is coupled to a circuit tuned to the crystal frequency. This tuned circuit is also coupled to an amplifier by means of which the beats between the crystal and the generator harmonics are amplified and transmitted to the phones.

(6) The amplifier consists of three radio stages and a detector.

Making a Run.

The frequency of the standard crystal is first determined approximately by means of a standard wavemeter. This is accurate to about 0.1 of 1 per cent.

A sheet of paper is placed on the drum and the generator started up. After the generator has warmed up for about 30 minutes the phones from the amplifier shown in Fig. 10 are plugged into the generator room. The operator, with one hand on the motor generator flywheel, varies the generator speed slightly from 500 cycles until some multiple or harmonic of it approaches the crystal frequency when beats are heard in the telephones. The operator now adjusts his speed until slow beats are heard. With the assistance of the other operator the Maxwell bridge is balanced, which brings the light beam to rest on the scale. This is done while the first operator is controlling the slight changes in generator speed by changing the pressure upon the flywheel.

The only function of the bridge is to indicate to the operator by the direction of the deflection of the beam of light whether fast beats in the phones represent an increase or a decrease in generator speed, so that he may quickly increase or decrease pressure on the flywheel to restore the generator speed to its proper value as indicated by very slow beats in the phones. The note in the phones is of the pitch of the generator fundamental frequency, which fact may be made clear as follows: when any harmonic of the generator frequency is at zero beat with the crystal, the harmonics next above and below this one give a beat note with the crystal. If the harmonic at 25,000

cycles is tuned to the crystal, then the one at 24,500 and at 25,500 give a 500-cycle note with the crystal. Pulses or beats in this note indicate how much the 25,000-cycle harmonic differs from the crystal frequency. If a few beats are heard on the fast side, the generator speed is reduced just enough to give the same number of beats on the slow side of the zero beat position. A careful operator may thus compensate for slight changes in beat frequency during a run.

The operator now controls the generator speed to give as slow a beat frequency as possible for a few minutes before starting to record the run. When he notes that the machine is stable enough, he closes the chronometer circuit and the

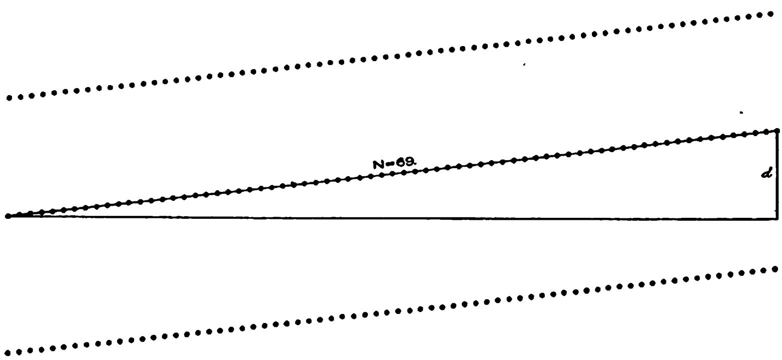

Fig. 11—Representation of Partial Chronograph Record Made in Calibrating a Crystal.

striker commences to put a record upon the paper mounted on the revolving drum. The time required to make a run varies from 8 to 15 minutes.

After the run the second operator may make a duplicate run upon the same harmonic, and then both operators make runs on another harmonic. No error due to the human element is observable in the results.

Obtaining the Generator Frequency.

Before removing the paper record from the drum a line is drawn on it parallel to the axis of the drum. The completed record will contain six parallel rows of dots since the drum makes one complete revolution in approximately six seconds, depending upon the frequency of the crystal. The striker puts

a dot on the paper every second, while the carriage upon which the striker rests advances approximately 1 millimeter per revolution of the drum. In Fig. 11 a partial representation of a record is shown. This record does not apply to the 25 kc. standard oscillator.

The method for computing the generator frequency is based on the fact that owing to the number of poles in the generator and the ratio of the reduction gear between the generator and the chronograph drum as previously mentioned, a rotation of the drum in exactly six seconds corresponds to a generator frequency of exactly 500 cycles. That is, if the six parallel rows of dots printed on the drum through the chronometer are parallel to the axis of the drum the generator is operating at 500 cycles.

If the slope of these rows of dots is up, the drum has not made a complete rotation in six seconds and therefore the generator speed is less than 500 cycles. A downward slope indicates more than one rotation of the drum in the six second interval or a generator speed greater than 500 cycles.

Assume for example that the rows of dots are parallel to the axis of the drum. Then in 100 rotations a point on the surface of the drum would have traveled 100 times the circumference of the drum, 100 x 51.9 cm. This corresponds to a 500-cycle generator speed. By a simple proportion the generator frequency for any slope of line can be determined thus:

$$CN : CN \pm d = 500 : X$$

where C = the circumference of drum
N = number of rotations of the drum
d = the perpendicular distance between the first and the last dot in a line
X = the generator frequency.

The d above is the difference between the distance actually traveled by a point on the circumference of the drum during a run, and that corresponding to the nearest number of complete revolutions of the drum. If the slope is up, d is negative; if down, the sign is positive.

Example

$C = 51.9$
$N = 100$
$d = 0.30$ cm. Slope is up. Therefore d is negative.

Generator frequency = 499.971 cycles per second.

From the measurement of the crystal frequency by the standard wavemeter, its value was found to be roughly 25 kc. Therefore, it must have been the fiftieth multiple of the generator fundamental that was held during the run at the crystal fundamental frequency.

The crystal-oscillator frequency is therefore fifty times the generator frequency or 24,998.6 cycles. If the generator speed is increased sufficiently, a point will be found around 510 cycles where the zero beat condition may be again located by means of the phones. This is obviously the forty-ninth harmonic of the generator adjusted to the crystal frequency. ($510 \times 49 =$ 25,000 approximately).

Accuracy of the Method.

The accuracy of a calibration is chiefly dependent on the constancy to which the generator frequency can be held. To assist in this, the motor-generator is operated on a bank of high capacity storage batteries kept in good condition and fully charged. The machine is allowed to warm up thoroughly before a run is attempted. The Maxwell bridge battery is also carefully checked for constant voltage, and the brushes and contactor segments in the bridge circuit are kept clean and properly adjusted. As previously stated, slight variations in the generator frequency can be compensated for by the operator, but if the run is noticeably poor it is so recorded, or discarded.

The error arising from the measurement of the line d is minimized by determining it for all six parallel lines in a record, and using the mean as the correct value for d. The resulting error does not average more than one part in 200,000. The combined errors then, on a good run, are well within 0.001 per cent. This degree of accuracy is possible only when the crystal is left undisturbed. The accuracy claimed for the standard crystal oscillator when used as a portable standard is only 0.002 per cent. No appreciable error is observable due to a probable error in the measurement of the drum circumference.

To give an idea of the correspondence of different calibrations a table of eight consecutive runs on a standard crystal is appended. It is, of course, understood that the last figure in the frequency column below is carried along merely to discriminate further between the measured values for different calibrations. It will

be noted that the probable error in the mean value is three parts in two and one-half million.

TABLE OF CALIBRATIONS

Date	Temp. Deg.	Frequency, cycles
8-15-27	38.45	24998.60
8-15-27	38.45	24998.75
8-16-27	38.45	24998.71
8-16-27	38.45	24998.73
8-29-27	38.55	24998.99
8-29-27	38.55	24998.82
9- 8-27	38.50	24998.71
9- 8-27	38.50	24998.63
Mean value	=	24998.74

Uses of the Standard Crystal-Controlled Oscillator

This standard oscillator is a portable device, the size being 8 x 14 x 22 1/2 inches. It is simple to use, there being no variables or tuning controls. It requires a plate potential of only ninety volts, and the heater may be operated on the 110-volt line. By its use, two hundred or more fixed frequencies equally spaced in kilocycles and accurate to about 0.001 per cent are made available. The permanence of the crystal calibration depends only on the permanence of the crystal itself.

This oscillator is used at the Naval Research Laboratory to standardize all quartz crystals manufactured there. Through the medium of a precision heterodyne frequency meter with a straight line frequency characteristic, a crystal of any frequency can be calibrated from the standard oscillator to about 0.002 per cent.

Furthermore, all heterodyne frequency meters for the Naval Service are calibrated directly against the harmonics of a standard oscillator. Readings may be taken as rapidly as the heterodyne frequency meter can be tuned to the successive harmonics of the crystal oscillator. For very accurate work, readings are taken on a beat note of 1000 cycles (as determined by a tuning fork) on both sides of the zero beat point.

The calibration of a heterodyne frequency meter can be rapidly checked to determine the error at any time, by taking readings on a few harmonics of the crystal. For this purpose a special type of crystal oscillator without temperature control has been furnished the Fleet under the name of the "Crystal-Controlled Calibrator," Navy Type No. SE 2907.

The standard oscillator may also be used for the intercomparison of frequency standards. One such comparison has already been made between the Navy's standard crystal-controlled oscillator and the Bell Telephone Laboratories' standard tuning-fork, to determine the relative constancy and the agreement in calibration of the two standards. For this test, the Navy oscillator was transported to New York City. A number of oscillographic records of the beat frequency between harmonics of the two standards taken at intervals over a period of three days indicated that there was no noticeable variation in the relative constancy of the two sources. The determination of frequency by the two standards differed by one part in 80,000 or 0.0012 per cent. This agreement is considered excellent when it is remembered that the Navy standard had been taken from Washington to New York for the test, and that entirely different methods were used to calibrate the two standards.

Since this comparison with the Bell Telephone Laboratories' standard was made, this Laboratory has built four crystal-controlled oscillators similar to the one described in this paper, as a part of a program for the establishment of a national secondary frequency standard decided upon at a conference in Washington in February of last year. This conference was attended by representatives of the various interested government departments, and of a number of the radio manufacturing companies.

It is hoped that, after a determination of the frequency of these oscillators at several other laboratories has been made, and an average result arrived at, they serve as useful secondary standards of frequency for the government and the radio industry of the country.

A TRANSMITTER MODULATING DEVICE FOR THE STUDY OF THE KENNELLY-HEAVISIDE LAYER BY THE ECHO METHOD*

By

M. A. Tuve and O. Dahl

(Department of Terrestrial Magnetism, Carnegie Institution of Washington, Washington, D. C.)

Summary—The importance is emphasized of sending out "peaks" of very short duration and proper spacing for the study of the radio reflections from the ionized layer in the upper atmosphere by the echo method, and objections to modified alternating-current modulation are pointed out. A method is outlined for modulating the transmitter, based on the sudden pulses of plate-current which occur in an unbalanced multivibrator circuit. The application of this method to the transmitter is described and its effectiveness illustrated by typical recorded signals.

In the study of the Kennelly-Heaviside layer by the *echo* method[1] a radio transmitter is modulated to emit a set of spaced "peaks" or "humps" of audio frequency, and the resulting "peaks" at the receiver are recorded by an oscillograph. If the transmitted peaks are of sufficiently short duration, single, double, or even multiple peaks are recorded at the receiver, corresponding (under proper conditions) to the ground wave and to zero, one or more "sky waves" arriving at the receiver via the reflecting layer in the upper atmosphere. In the previous work referred to above[2] modulation of the transmitter was accomplished by superposing an alternating e.m.f. of 500 to 1,500 cycles on a high negative bias applied to the grid of the intermediate amplifier in the transmitter. Under these conditions, power is radiated during a part of each positive half cycle of the alternating e.m.f., the time between successive peaks being only slightly greater than their duration or "on-time." Excessively high voltages would be required to make the "on-time" very much less than one half-cycle. The resulting duration of the peaks at 500 cycles was too long to give complete resolution at the receiver for small heights of the layer, and always failed to resolve completely the multiple "reflected peaks" which were

* Original Manuscript Received by the Institute, March 2, 1928.

[1] G. Breit and M. A. Tuve. *Phys. Rev.*, 28, pp. 554–575, 1926. O. Dahl and L. A. Gebhard. Proc. Inst. Radio Engineers, 16, No. 3, March 1928.

[2] Since the work described in this paper was done, a publication by R. A. Heising has appeared (Proc. Inst. Radio Engineers, 16, No. 1, Jan. 1928) containing an account of the work he has done using a similar method. His *peaks* were of duration 0.001 and 0.0016 second, and he did not obtain complete resolution except under unusual conditions.

received at times. To eliminate the possibility that false peaks may be produced by interference effects, it is necessary that the waves shall arrive separately at the receiver, giving peaks which are completely separated on the base-line of the oscillograph record. An additional objection to this method of modulation was that because of the uniform and short spacing between successive peaks there was ambiguity as to whether a given "reflected peak" was due to the "ground peak" immediately preceding it or to an earlier one. Rapid interruption or "keying" of the transmitter eliminated this ambiguity, provided the end of a given signal happened to be recorded on the fast-moving film at the receiver. These objections pointed clearly to the need for a shorter "on-time" and a longer interval between peaks.

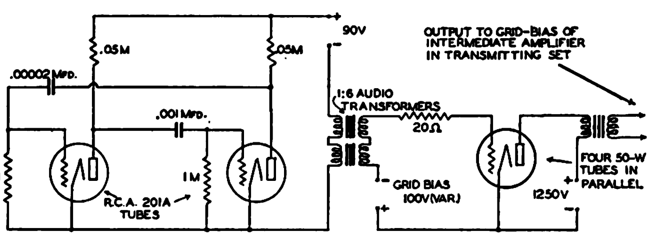

Fig. 1—Diagram of Connections. The Resistance of 20 Ohms in Grid Circuit of 50-Watt Tubes is Put Very Close to the Grids to Prevent Ultra High-Frequency Oscillations.

The use of 1,500-cycle modulation gave peaks of shorter duration, but they were also more closely spaced, and this resulted for the most part in still greater confusion.

When a "multivibrator" circuit as described by Abraham and Bloch[3] is used in a very unbalanced condition, an oscillograph shows the existence of very sharp and widely spaced pulses of current superposed on the otherwise steady total plate-current of the two tubes. The duration and frequency of these pulses may be controlled readily by varying the resistances and capacities interconnected between the plates and grids of the two tubes. Using 201-A tubes with circuit resistances of 0.1 and 1 megohm and capacities of 0.0002 and 0.001 μfd. (see Fig. 1), the "frequency" of the pulses (their spacing is somewhat irregular) is about 300 per second and their duration perhaps 0.0002 second. This immediately suggests a means of modulating a transmitter for the radio "echo" experiments. A transformer in the plate-supply circuit of these tubes converts a current pulse

into two voltage impulses (plus and minus), each of shorter duration than the current pulse. Applied to the grid of an amplifier with a suitably high negative bias, only the voltage impulse in the positive direction affects the tube, resulting in similar pulses of even shorter duration (due to the grid bias) in its plate current. Again transforming these, a sufficient voltage may be obtained to modulate a radio transmitter (see Fig. 1).

Using 201–A tubes for the multivibrator and one 50-watt tube as an amplifier, a transformer in the plate circuit of the 50-watt tube gave peaks of 1,200 volts and of extremely short duration (less than 1/4,000 second). The amplifier gives "clean" pulses of short duration in its plate current only if the input transformer to its grid is so connected that the initial rise of the multivibrator current for one peak superposes a negative e.m.f. on the negative amplifier grid bias, and not vice versa. Thus the amplifier pulses are due to the rate of change of the current at the peak of the pulse, and not at its beginning. This multivibrator and amplifier set was applied to the 20-kilowatt, 4,015-kilocycle transmitter at the Naval Research Laboratory, Anacostia, D. C., which has been used previously in the reflection experiments.[1] In this transmitter two 250-watt tubes in parallel are used as the intermediate amplifier between the crystal-controlled master oscillator and the 20-kilowatt tube, and a change of several hundred volts on the grids of these tubes is sufficient to modulate the transmitter from zero to full power. It was consequently expected that the above multivibrator set would serve very well for direct grid modulation on these tubes. However, when it was connected, the transmitter was modulated by each peak to only a fraction of its full-power emission, as shown by comparison of the received signals using 500-cycle modulation.

When the transmitter is radiating full power, the bias on the grids of the 250-watt tubes is about 100 volts negative. However, the radio-frequency grid voltage superposed on this bias is sufficient to make the grids positive during a fraction of each radio-frequency cycle, resulting in instantaneous values of grid current which may be of considerable magnitude, and it is clear that the device (the modulator) supplying the grid bias must maintain its voltage when this current is drawn, i.e., must supply power to the grids of these tubes. When four 50-watt tubes in parallel were substituted for the single 50-watt amplifier tube,

[3] H. Abraham and E. Bloch. *Annales de Physique* 9, 12, 1919, p. 237.

the resulting output was amply sufficient to modulate the transmitter to full power. Actually the peak value of the power using this modulation is somewhat greater than the usual full power for continuous output.

The resulting modulation of the transmitter is almost ideal for the study of the Kennelly-Heaviside layer by the echo

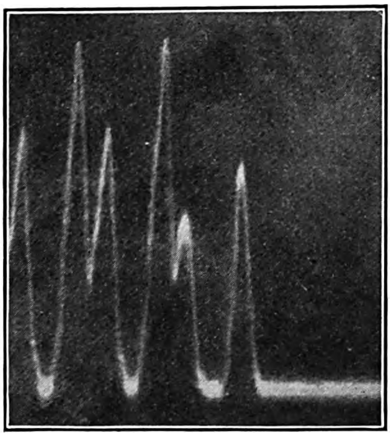

Fig. 2a—Reflections from a 146-mile Layer using 500-cycle Modulation.

method. The "frequency" of the peaks is readily varied, their duration is extremely short (less than 1/4,000 second), and their spacing is somewhat non-uniform. The latter fact is of great usefulness in the identification of the separate peaks due to the

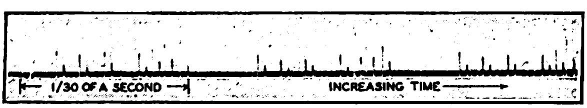

Fig. 2b—Reflections from a 137-mile Layer Using Multivibrator Modulation.

"ground" and "sky" waves. The peak due to the "ground wave" is completely separated from the "reflected peaks" even for very low heights of the layer, eliminating any possibility of interference effects. By reason of the same possibility of interference effects, the brief "on-time" is of great importance in the study of the apparent multiple reflections which occur at times. With this modulation, these multiple peaks are usually completely resolved. Fig. 2a shows reflection from a layer 146 miles high

using 500-cycle modulation. Fig. 2b shows reflection from a layer 137 miles high using the multivibrator modulation. The measured apparent width of the peaks is about 1/4,000 second. However, records made with incomplete damping of the oscillograph indicate that this is only an upper limit set by the period of the oscillograph itself. The actual duration of the peaks may be considerably less. The peaks emitted with the multivibrator modulation are not of uniform amplitude, as may be noted. This provides a useful check on the identification of the various peaks, and also on possible distortion in the receiving apparatus.

The writers are indebted to Dr. G. Breit for suggesting the method, and to Messrs. G. Breit, M. H. Schrenk, and L. A. Gebhard for assistance and cooperation in its application.

A COMPENSATED ELECTRON-TUBE VOLTMETER

By

H. M. Turner

(Yale University, New Haven, Connecticut)

A SOURCE of error sometimes encountered in the use of electron-tube voltmeters is that due to unavoidable changes in filament current especially when the filament is purposely operated well below the saturation value in order to stabilize the calibration over long periods of time. In Fig. 1 is

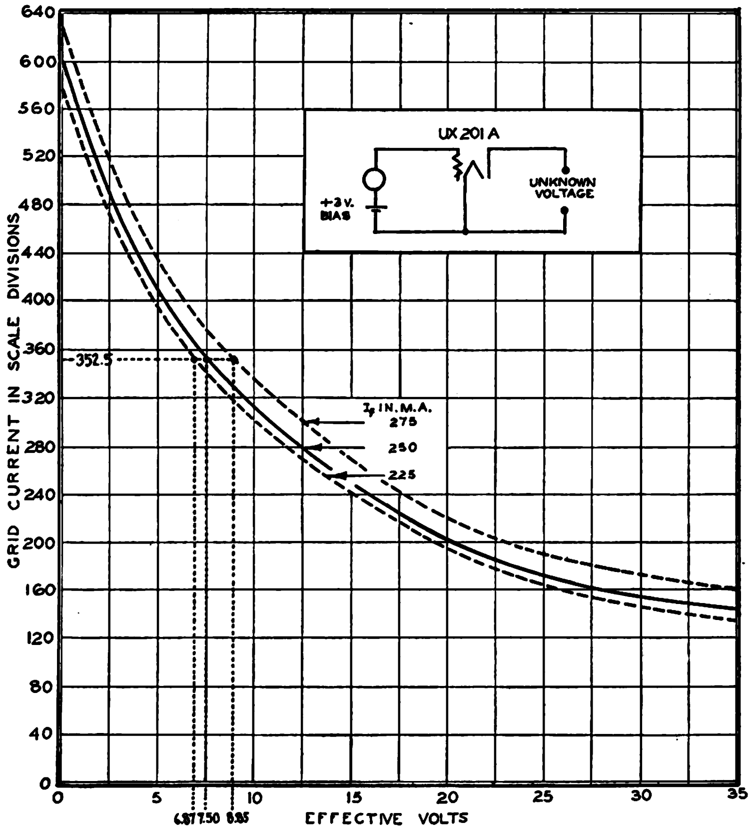

Fig. 1—Uncompensated Electron Tube Voltmeter; Grid Current Method.

shown a circuit arrangement for the "Grid Current Method" together with the calibration for normal filament current of 0.25 amperes. Should the normal calibration be used when the fila-

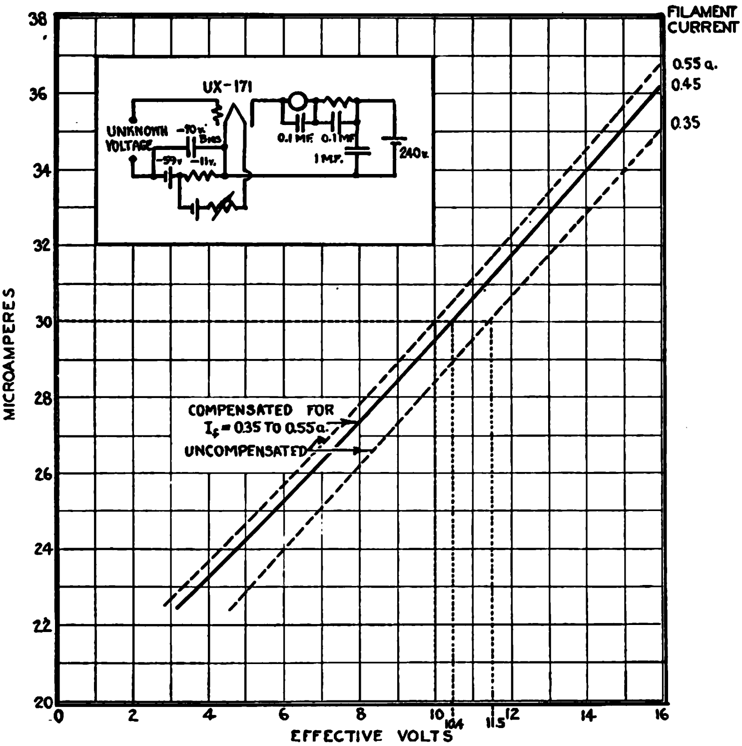

Fig. 3—Compensated Electron Tube Voltmeter; Plate Current Method.

Current Method;" the calibration curve (solid) is good between the limits 0.35 and 0.5 amperes. The dotted curves show that for the same filament current limits the error for the uncompensated method would be 10.6 and 4 per cent respectively.

NOTE ON THE EFFECTIVE HEATING OF CODE TRANSMITTERS*

By

Frederick Emmons Terman

(Department of Electrical Engineering, Stanford University, California)

Summary—By considering the frequency of letters and words in the English language it is found that in slow speed transmission of code using the theoretical spacing, the transmitter is in operation only 46.5 per cent of the time. A transmitter can accordingly be given a power rating for code that is 2.15 times the rating of the same equipment when operating continuously, on the basis of the same operating temperature. At high speeds of transmission the heating is slightly less than at slow speeds. The desirability of designing large transmitters to take advantage of the full code rating possible during periods of weak signals or heavy static is brought out.

The power rating of radio transmitters and radio transmitting apparatus is determined either partially or entirely by the allowable temperature rise during operation. In most types of code transmitters the full power is on only during the part of the time the key is depressed. In rating transmitters and transmitting equipment to be used for code purposes it is accordingly important to know the average rate of heating of the set over a long period of code operation compared with the rate heating that would take place with full power continuously on.

The average rate of heating depends somewhat upon the speed of transmission. At slow speeds the transmitted code impulses are substantially rectangular, as shown in Fig. 1a, but at high speeds the time required for the signal characters to reach approximately full value is an appreciable part of the dot time interval, and the current takes a corresponding time to die out after the key is opened.

The rise and fall of transmitted current follows an exponential law, giving high speed dot and dash characters the shape shown in Fig. 1b.

Taking into account the number of dots and dashes required for each letter, the frequency with which each letter is used in the English language, etc., it is found that a transmitter in the act of sending code with the theoretical spacing is in operation only 46.5 per cent of the time. With slow speed transmission

* Original Manuscript Received by the Institute, April 19, 1928.

the average rate of heating is therefore 46.5 per cent of the rate of heating with full power on. In high speed transmission the average rate of heating is a little less. With an exponential law for the rise and fall of transmitted current, *assuming that the instantaneous heating is proportioned to the square of transmitted instantaneous current,* and neglecting the heating of the small current that remains one dot period after the key has been opened, the following table holds:

TABLE I

HEATING OF CODE TRANSMITTERS

Completeness of current rise in one dot interval	Average rate of heating with code	Power rating for code / Rating for continuous use
1.00	0.465	2.15
0.98	0.364	2.75
0.95	0.347	2.88
0.90	0.312	3.20
0.80	0.278	3.60

The first column introduces the transmission speed in terms of the nearness with which the current reaches the steady state

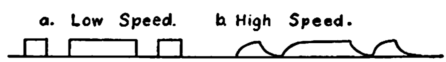

Fig.1—Shape of Transmitted Impulses at High and Low Speeds of Transmission.

value during the sending of a dot. A value of 1.00 represents slow speed transmission shown in Fig. 1a. The second column gives the average rate of heating in terms of the rate of heating when full power is continuously on. The third column is the reciprocal of the second column, and gives the ratio of power rating for code to power rating for continuous transmission for equal average heating (assuming heating is proportional to current squared).

When the thermal capacity of the equipment is large, as in the case of high power transmitters, the operating temperature depends upon the average rate of heating over an appreciable time interval. The third column of Table I shows that such transmitters can safely transmit code at over twice the power output permissible when energy is continuously radiated.

A code transmitter adjusted to give the maximum permissible output when sending code would be seriously overheated if allowed to radiate this energy continuously for a short while Protective relays to prevent such overheating are essential, as

continuous transmission is desirable for tuning-in and testing, and forgetful operators will hold the key down. Thermally-controlled relays could be used to lower the power output to the value safe for continuous operation in the event danger of over-heating was imminent.

The extent to which the full possibilities of the power rating for code transmission can be taken advantage of with safety depends upon the circumstances. In a large, well-supervised land station, with protective thermally-operated relays, it should be possible to realize all the benefits of the full code rating. Where it is not thought advisable to go to the limit, the service rendered by the station could be very greatly improved by designing the power supply and other equipment to make possible the use of the code power rating during periods of heavy static or weak signals.

The writer is indebted to Dr. J. E. Coover of the Stanford University Psychology Department for information giving the frequency with which different letters are used in English. Computations based on Dr. Coover's count show that 229 words, 1517 dots, and 1008 dashes are transmitted when sending 1000 letters.

FOUR-ELEMENT TUBE CHARACTERISTICS AS AFFECTING EFFICIENCY

By

David C. Prince

(Research Laboratory, General Electric Company, Schenectady, New York)

Summary—*The variation in the ratio of grid and plate current of three-electrode vacuum tubes aroused considerable curiosity and the tests reported in this paper were undertaken in an endeavor to ascertain the laws of current division. It was found that in a tube having symmetrical electrodes, that is, straight wire filament, concentric cylindrical anode and cylindrical grid, made up of wires parallel to the axis, the ratio of grid and plate current was a function of the tube geometry and quite different from that usually found in commercial design. The ratio of grid to plate current in such a tube exceeds the ratio of projected grid area by only a small amount, easily accounted for by variations in the electric field around the grid wires. The considerable departures from this ratio in commercial tubes appear to be due to a combination of secondary emission from the tube anode and unsymmetrical arrangements of grid wires and supports.*

Object

THIS work was undertaken because of the apparent lack of any logical proved explanation for the amount of current collected by the grid of a three-element tube of standard design when the grid and plate are at nearly the same potential.

At first glance it might appear that a tube having two electrodes, a cathode and a grid shaped anode should have nearly the same impedance characteristics as one having a solid plate. The following qualitative argument should show that this is not so. The current for any potential difference less than temperature saturation values is limited by space charge, that is the cumulative electrostatic field due to the electrons in the space between the electrodes. Individual electrons, when emitted, are acted upon by the electrostatic field of the anode and move in that direction since their initial velocities are relatively small. As the electrons approach the anode their distribution is more or less uniform. Due to their mass they cannot follow the lines of force to the anode wires and, in departing from those lines, some will fail to strike the anode. Those electrons which pass through accumulate in the outside space until their space charge in that zone is sufficient to drive some of them back toward the anode when a

* Original Manuscript Received by the Institute, February 1, 1928.

portion will again pass through. Those electrons which pass a second time add to the charge and therefore reduce the number which can be emitted in unit time. This is equivalent to an increase in impedance.

For a cylindrical tube as long as the motion of the electrons is radial the number of electrons striking the anode on the outward trip should be proportional to the projected area of the anode wires. On the return trip the same fraction in addition should strike the wires. The large additional numbers which do strike must then be due to turbulence.

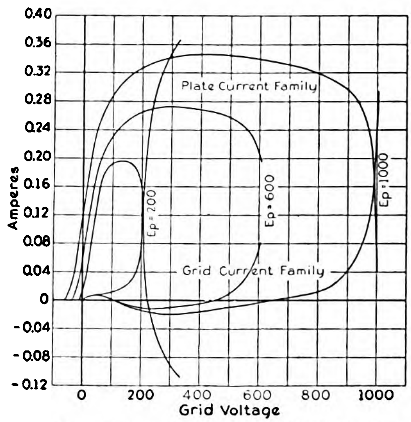

Fig. 1—Characteristics of a 250-Watt Pliotron.

Electrons which have been drawn into space by a positively charged grid structure may be collected by any electrode which lies in their path and is more positive than the cathode. If a large percentage of all the emitted electrons could be collected by an anode at low potential after having been drawn away from the cathode by a highly charged electrode, the total loss of energy in the tube due to space charge could be made very low.

This report covers the making of tests and calculations in a search for the requirements of such a low loss tube.

Work Performed

In a three-electrode tube a large number of electrons should pass through a positively charged grid even with the plate at cathode or ground potential. With the plate at positive potentials considerable current should flow. That this is not the case

is strikingly shown by Fig. 1 which is the characteristic of a 250-watt pliotron. At 300 volts grid and 200 volts plate, the plate not only does not collect electrons, but actually emits them. This is unmistakable evidence that secondary emission from the plate contributes to the supply of electrons which may reach the grid through turbulence. The problem in hand can, therefore, not be studied effectively unless this phenomenon can be excluded. To eliminate secondary emission, recourse was had to a fourth electrode. This electrode was in the form of a second grid placed between the usual grid and the plate. By maintaining this grid below plate potential, a negative gradient is main-

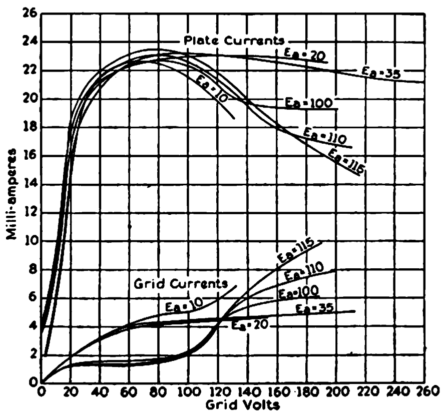

Fig. 2—Characteristics of Tube A with 125 Volts on Plate. E_a is the Anode or Outer Grid Potential.

tained at the surface of the plate, tending to drive secondary electrons back to the plate even though emitted with considerable velocity.

The first of these tubes to be tested was a small receiving type with oval elements designated Tube *A*. The observed characteristics are shown in Fig. 2. It will be observed that the tendency of the plate to lose electrons to the grid is checked by lowering the potentials on the anode or outer grid. With anode volts E_a equal to 10, the plate again loses electrons due to some cause not obvious but probably connected with turbulence effects.

Since the tests of Tube *A* showed that the loss of electrons by the plate and gain by the grid could be positively reduced by the fourth electrode, a larger test sample, Tube *B*, was made

up from parts of a 1 kw. pliotron. The inner grid was, however, made with coarser mesh than the standard tube in order that the combined effect of the two grids might be nearly the same as

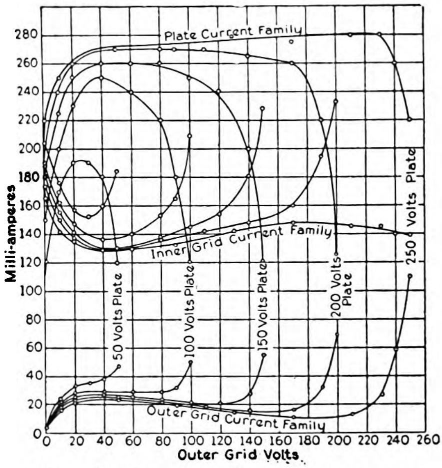

Fig. 3—Characteristics of Tube B with 250 Volts on Inner Grid and 15 Amperes Filament. Principal Dimensions: plate, 3¼ in. long by 1½ in. in diam.; filament, 6⅛ in. of 18 mil. wire (V-shaped); outer grid 1¼ in. in diam. 13 turns per inch of 10 mil. wire; inner grid, ¾ in. in diam. 13 turns per inch of 10 mil. wire.

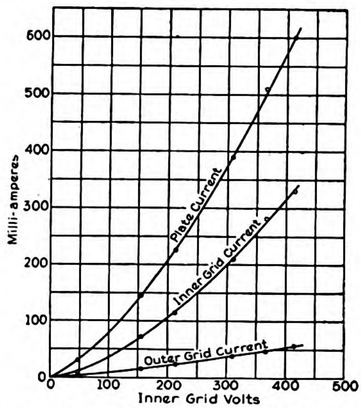

Fig. 4—Characteristics of Tube B with 150 Volts on Plate, 50 Volts on Outer Grid, and 16.5 Amperes Filament.

that of the single grid in the standard tube. The characteristics of this tube are shown in Figs. 3 and 4.

From Fig. 3 it appears that about 150 volts on the plate are required to get substantially full current and that, for this po-

tential, 50 volts on the outer grid give minimum inner grid current. Maintaining these conditions, Fig. 4 gives the variation of current with inner grid voltage. It is apparent that throughout the range of this test the currents to all three positively charged electrodes are substantially proportional and that the plate current is highest even though the inner grid is most positive of any of the elements. Nevertheless the inner grid current shown is out of all proportion to the grid area.

It seemed a fair assumption that the plate secondary emission in Tube B is very small under the conditions used. The large grid current was therefore attributed to lack of symmetry and resulting turbulence. It was decided to construct a tube having the greatest possible symmetry, especially in the inner grid. A sketch of this tube is shown in Fig. 5. The filament, inner grid and plate of this tube are very perfect geometrically. The outer grid is made with longitudinal supports and helical wires, so that its symmetry is much less perfect, but the nature of this structure seemed relatively less important.

Characteristics of this tube, which has been designated Tube C, are shown in Figs. 6, 7, 8, and 9. A discussion of these observations follows the discussion of the theory.

Theory

Assume a cylindrical tube having a cross section such as shown in Fig. 10. Referring to this figure and Fig. 11, which is an enlarged portion of Fig. 10, let:

r = radius of grid

r_1 = radius of cylinder which, if substituted for the grid, would produce the same field at the filament

$r_2 = r - c$ = radius of an imaginary polygon

s = distance between grid wires

ρ = radius of grid wires

p = velocity acquired by an electron in passing from the filament to the polygon

v = tangential component of p referred to nearest grid wire (Fig. 11)

u = radial component of p referred to nearest grid wire

$u_0^2 = p_1^2 - p^2$ = increase in square of velocity between polygon and grid wire

p_1 = velocity of an electron reaching the grid

$$v = p \sin \alpha$$

$$u = p \cos \alpha$$

$$u^2 + v^2 = p^2$$

d = distance from center line joining filament and grid wire at which an electron can pass the boundaries of the polygon and just come tangent to the grid wire

An examination of the static field produced by a grid shows the irregularities produced by the individual grid wires practically

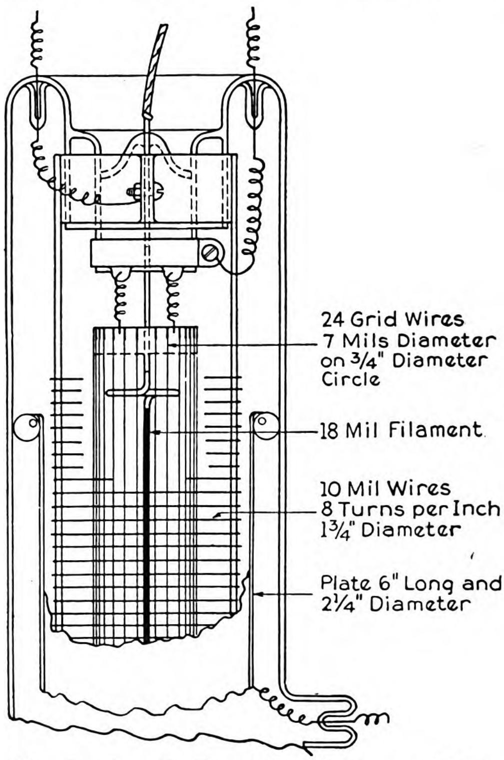

Fig. 5—Tube C—Four-Element Vacuum Tube.

disappear at a certain distance from the grid. Suppose an imaginary polygon to be drawn inside the grid at this distance. It seems reasonable to assume that the motion of an electron may be approximated closely by assuming that it moves inside the polygon in response to a field radial with respect to the filament, and outside the polygon in response to a field radial with respect

to the grid wires. Since the electrons which might strike one grid wire are diverging very slowly, the approximation is very little affected by assuming that the electron paths are normal to the sides of the polygon as they cross it.

Under these assumptions

$$\alpha = \tan^{-1}\frac{d}{c}$$

The radius r_1 is determined by the formula

$$\log_e\frac{r_1}{r} = \frac{s}{2\pi r}\log_e\frac{s}{2\pi\rho} \tag{1}$$

This is derived in the conventional way, assuming the grid wires small compared with their spacing and neglecting the presence of the plate.

Inside the polygon the electrons acquire a velocity such that

$$p^2 = Kr_2^{2/3} = K(r-c)^{2/3} \tag{2}$$

The potential of the grid is the same as that of the equivalent cylinder of radius r_1 so that the velocity of an electron reaching the grid is

$$p_1^2 = Kr_1^{2/3} \tag{3}$$

The increase in the square of the velocity between the polygon and the grid wires is then

$$u_0^2 = p_1^2 - p^2 = K\left[r_1^{2/3} - (r-c)^{2/3}\right] \tag{4}$$

Dr. Hull[1] has developed the following expression for the condition under which an electron emitted by a cylindrical cathode of radius R with radial velocity u and tangential velocity v will just come tangent to an internal cylindrical anode of radius ρ

$$\frac{2e}{m}V = \left(\frac{He}{2m}\right)^2(\rho^2-R^2)\left(1-\frac{R^2}{\rho^2}\right)+\frac{He}{m}\,\frac{Rv}{\rho^2}(\rho^2-R^2)$$
$$-\frac{v^2}{\rho^2}(\rho^2-R^2)-u^2 \tag{5}$$

If there is no magnetic field present (5) reduces to

$$\frac{2e}{m}V+u^2 = -\frac{v^2}{\rho^2}(\rho^2-R^2) \tag{6}$$

[1] A. W. Hull, "The Effect of a Uniform Magnetic Field on the Motions of Electrons between Coaxial Cylinders," *Physical Review*, 18, No. 1, July, 1921.

Since $\frac{2e}{m}V$ is the gain in the square of the velocity in passing between cathode and anode and $R^2=c^2$, (6) may be rewritten

$$\frac{u_0^2+u^2}{v^2}=\frac{c^2-\rho^2}{\rho^2} \tag{7}$$

Equation (7) then defines the condition under which an electron will just reach the grid wire. Substituting values

$$\frac{u_0^2+p^2\cos^2\alpha}{p^2\sin^2\alpha}=\frac{c^2-\rho^2}{\rho^2} \tag{8}$$

$$\frac{u_0^2+p^2}{p^2\sin^2\alpha}=1+\frac{c^2-\rho^2}{\rho^2}=\left(\frac{c}{\rho}\right)^2 \tag{9}$$

$$\sin\alpha=\frac{d}{\sqrt{d^2+c^2}} \text{ and } \frac{1}{\sin^2\alpha}=\frac{d^2+c^2}{d^2}=1+\left(\frac{c}{d}\right)^2 \tag{10}$$

$$u_0^2+p^2=p_1^2 \tag{11}$$

the final velocity of the electron

$$\text{therefore}\left(\frac{c}{d}\right)^2=\frac{p^2}{p_1^2}\left(\frac{c}{\rho}\right)^2-1\cdot \tag{12}$$

From (12) d can be determined, and it represents the distance from the center line joining filament and grid wire at which an electron, crossing the boundary of the polygon, will just come tangent to the grid wire. Electrons passing outside the distance d will not strike the grid wire, while electrons passing inside this distance will strike it.

Since the grid wires are separated by a distance s, the length of one side of the polygon is $s\frac{r_2}{r}$. Electrons passing either side of the center line within the distance d will strike the grid. The proportion of electrons which strike the grid is then the ratio of these dimensions, that is

$$\frac{i_g}{i}=\frac{2dr}{sr_2} \tag{13}$$

where i_g is grid current and i total emission.

For the Tube C under investigation

$\rho=0.0035$, $r=0.375$, $s=0.098$, $r_1=0.4$ from (1), $p_1^2/K=0.543$

Let $c/s=$	0.6	0.8	1.0	1.2
c	0.0588	0.0784	0.098	0.1176
$r_2=r-c$	0.316	0.297	0.277	0.257
$p^2/K=(r_2)^{2/3}$	0.464	0.445	0.425	0.405
$(c/\rho)^2=$	282	500	780	1120
$(c/d)^2=\dfrac{p^2}{p_1^2}(c/\rho)^2-1$	240	410	610	835
d/c	0.00645	0.00493	0.00405	0.00346
d	0.00379	0.00386	0.00397	0.00407
$i_g/i=\dfrac{2d}{s}\times\dfrac{r}{r_2}$	0.0917	0.0995	0.110	0.121

The relation between c/s and i_g/i is shown in Fig. 12.

From Fig. 7, for 250 volts on the inner grid, the emission is made up as follows:

Plate current	1.21 Amperes
Outer grid	0.13
Inner grid	0.15
Total Emission	1.49 Amperes

$$\frac{i_g}{i}=\frac{0.15}{1.49}=0.101$$

Projected area of grid $\dfrac{2\rho}{s}=0.071$

The grid current can thus be determined for this particular tube by assuming $c=0.83s$. The foregoing method of determining grid current is obviously empirical and the results of a test of one tube cannot be considered as quantitatively conclusive. However, the results seem to verify the general theory as conclusively as a single experiment can be expected to do. The main purpose of the investigation has thus been accomplished.

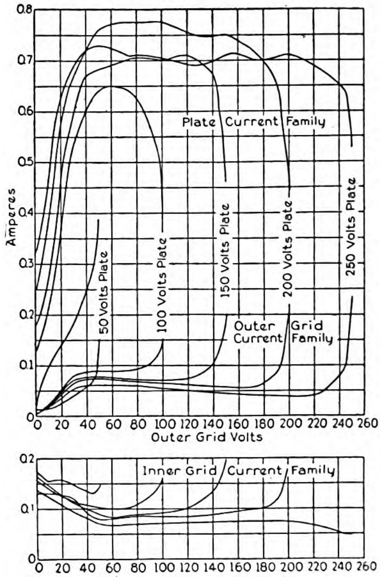

Fig. 6—Characteristics of Tube C with 250 Volts on the Inner Grid and 14.8 Amperes Filament Current.

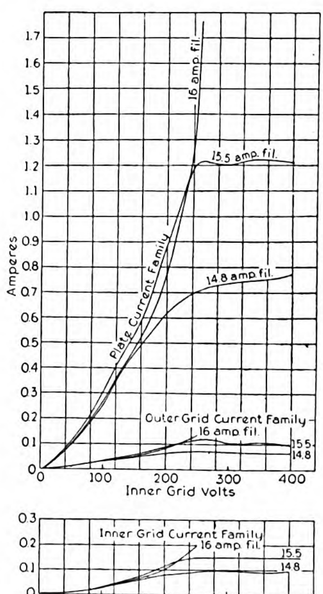

Fig. 7—Characteristics of Tube C with 150 Volts on Plate and 75 Volts on Outer Grid.

It is possible to make other observations of interest from a tube of this arrangement. In the foregoing theory, it has been assumed that all electrons passed in the direction from filament to plate. Suppose the plate to be at zero potential so that it does not receive the electrons. They then return and some strike the grid on the second passage. Let the electrons emitted be a, then if $0.9a$ pass the grid and, if none reach the plate, $0.9a$ return and $0.81a$ pass again into the zone between filament and grid. The number striking the grid is $0.19a$. The total electrons

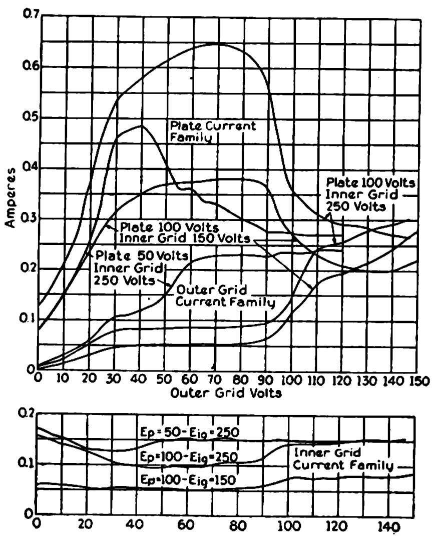

Fig. 8—Characteristics of Tube C with 14.8 Amperes Filament Current.

in the space determine the space charge current so that, if electrons were not passing both ways, the same total number would be emitted. Electrons in filament-grid zone are $1.81a$, and the electrons striking the grid are therefore $\frac{0.19}{1.81}=0.105$ as compared with 0.100 which would reach the grid were there voltage on the plate. The grid current should, therefore, be 5 per cent higher when there is no voltage on the plate than when there is a positive plate voltage. Fig. 9 shows that this condition probably exists within the limits of experimental error, although the difference might be somewhat greater than 5 per cent due to

tangential velocities acquired by the electrons which pass near to grid wires.

Dr. Langmuir[2] has derived relations giving the effect of cathode diameter on space charge. These relations are shown graphically in Fig. 13. These curves apply whether the cathode be inside or outside the anode, providing both are concentric cylinders. When electrons pass from the axial filament through the grid, they immediately begin to produce a space charge outside the grid and they will, therefore, decelerate and lose all

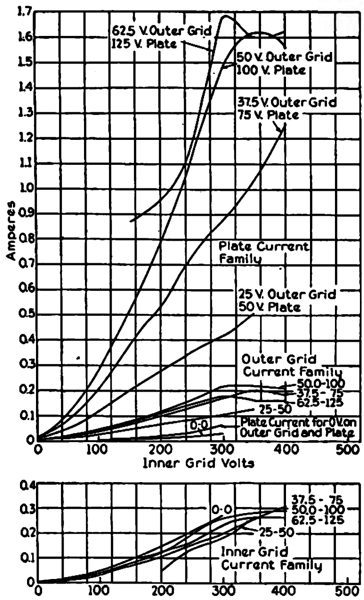

Fig. 9—Characteristics of Tube C with 16.0 Amperes Filament Current.

radial velocity when they have reached a certain distance. Whether the electrons move toward or away from the anode is immaterial, since space charge is a function of the number present and their distribution. The number passing any radius per second is the same and is equal to the current. The velocity distribution is the same since the velocities are zero at the cathode in either case and are equal to a constant times the square root of the voltage at any other radius.

[2] Irving Langmuir, "The Effect of Space Charge and Residual Gases on Thermionic Currents in High Vacuum," *Physical Review*, 2, 450-86 1913.

If the electrons are not collected when they come to zero radial velocity, they return again toward the grid and pass through it. The sum of all electrons inside the grid is determined by Fig. 13 for $r<a$; the sum of all electrons outside must be different only by those which strike the grid. A corresponding value of r/a is therefore obtained, giving a zone at which electrons come to rest. This zone then becomes a virtual cathode.

In making an approximate application to the sample tube, assume that the grid has the same effect as a cylinder of the same diameter. For the zone from filament to grid

$$\frac{r}{a}=\frac{0.375}{0.009}=42 \text{ so that } \beta^2=1 \text{ and}$$

$$i=14.65\times10^{-6}\frac{E^{3/2}}{r}$$

For the outer zone with electrons passing one way only

$$0.9i=14.65\times10^{-6}\frac{E^{3/2}}{r\beta^2}$$

Dividing $$\frac{i}{0.9i}=\frac{14.65\times10^{-6}\dfrac{E^{3/2}}{r}}{14.65\times10^{-6}\dfrac{E^{3/2}}{r\beta^2}}=\beta^2=1.11$$

for which $\dfrac{a}{r}=2.18$ or $a=0.818$ in. For the case where the electrons pass through the grid but are not collected by the anode and so return, current in the outer zone is

$$\frac{180}{181}i=14.65\times10^{-6}\frac{E^{3/2}}{r\beta^2}$$

from which $\beta^2=1,\qquad \dfrac{a}{r}=2.1,\qquad a=0.788$ in.

The radius of the outer grid is 0.875 in. We should expect then the equivalent of a three-element tube having a plate of radius 1.125 in., grid of 8 turns per inch 0.01 in. wire with 0.875 in. radius, the cathode varying between 0.788 in. and 0.818 in. radius.

The larger cathode radius would apply to the case where current is being drawn by the plate. However, it seems reasonable to suppose that those electrons which pass close to grid wires will have some of their radial velocity converted into tangential velocity. These electrons would not travel so far from the grid

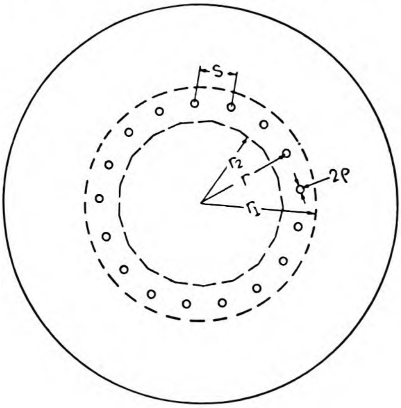

Fig. 10

as those which pass midway between grid wires. This effect would tend to make smaller the effective cathode radius as current draughts increase. Experimental data are not yet sufficient on this point. The apparent effective cathode-grid spacing, from a large number of test points, varies from 0.145 in. to 0.182 in.,

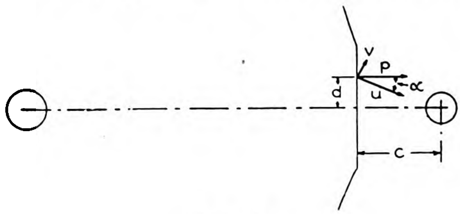

Fig. 11

whereas the theoretical difference is from 0.057 in. to 0.087 in. neglecting the path curvature effect just referred to.

An idea can be obtained from Fig. 6 regarding emission velocities of secondary electrons from molybdenum. The amplification constant of the outer grid referred to the inner grid as anode should be of the order of 25[3] so that the field due to the

[3] F. B. Vogdes and F. R. Elder, "Amplification Constant for Three-element Tubes," *Physical Review*, 21, pp. 683-689, No. 6, Dec. 1924.

inner grid at the plate is small compared with that due to the outer grid. Considering the outer grid only, the electrons appear to begin to leave the plate against a potential of about fifty volts negative on the outer grid.

Theoretical Conclusions

(1) With a symmetrical grid of the type used, the grid current is always comparatively small. As far as can be told with tests of only one tube, the number of electrons to strike the grid can be determined by assuming that the electrons enter a field which is radial with respect to the grid wires when they

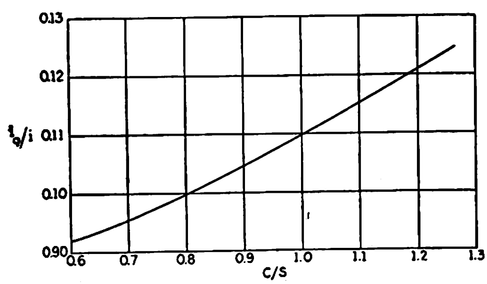

Fig. 12—Variation in Theoretical Grid-Plate Current Ratio with Change in Assumed Equivalent Static Diagram. Projected Grid Area 0.071.

cross a line normal to their paths and removed from the grid wires by a distance equal to 0.83 times the grid wire spacing.

(2) If the plate is protected from emitting secondary electrons, it does not lose current to the grid even though the latter is the most positive element in the tube. The plate voltage current characteristic, under these conditions, corresponds to a tube having a cathode considerably larger than the inner grid. The law covering the exact diameter of the virtual cathode has not yet been established.

(3) To prevent loss of secondary electrons to the grid, a potential on the outer grid, about fifty volts lower than the plate, is required with the tube under test.

Comparative Conclusions

The practical value of the four-element tube with symmetrical grid lies in the possibility of largely reducing space charge losses without recourse to gas effects, which ordinarily cannot be taken advantage of except at low voltage. Fig. 14 shows the watts per

ampere or equivalent volts drop for the various values of outer grid and plate drop. The figures include watts loss in plate and both grids. For comparative purposes, the losses are plotted for a kenotron having the same plate radius (1.125 in.) and for a kenotron having approximately the same loss. The radius of such a kenotron would be 0.125 in. Such a kenotron would not only be extremely hard to build with large current capacity, but the internal static stresses would probably be prohibitive. The loss per square inch of anode area at 1.6 amperes is 6.22 watts as compared with 58 for the 0.125 in. radius kenotron, not including filament energy.

This four-element tube appears to combine an impedance several fold lower than is obtainable with an ordinary high-vacuum kenotron, and the grid-control and filament shielding of a high-voltage pliotron.

DETECTION WITH THE FOUR-ELECTRODE TUBE*

By

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Summary—*A mathematical analysis of plate rectification is presented. Results of this analysis are applied to the screen-grid tube of the type Cunnningham CX-322. Experimental data verifying the mathematical analysis as applied to the CX-322 are presented.*

The results show that the screen-grid detector tube under proper conditions will efficiently utilize the high radio-frequency voltage obtained with the screen-grid tubes used as radio-frequency amplifiers. The square law holds for large input voltages making it practical for the detector to work power tubes of the type Cunningham CX-371.

In an investigation of the various applications of the four-element tube of the screen-grid type in the field of broadcast reception, it was evident that the voltage amplification is far above that obtained with an equal number of three-element tubes, and usually higher than that required. When an attempt is made to reduce the number of tubes in the receiver several practical difficulties arise. It is convenient to summarize first the performance obtained with the tubes used, Cunningham type CX-322, before discussing this point in detail.

In service as a radio-frequency amplifier using a tuned output circuit of average good quality a voltage amplification of 25 per stage at 550 kc. and of 40 per stage at 1500 kc. is readily obtained. This amplification is very much higher than that obtainable with three-element tubes, with which an amplification of ten per stage is perhaps as high a value as is practical to use because of selectivity requirements. In the above case the selectivity using the CX-322 is slightly better than the three-element tube. The simplest and most practical method of coupling radio-frequency amplifiers is the usual method of using a single tuned curcuit between tubes. With the conventional circuits at least three tuned stages will be required to obtain sufficient selectivity.

A high audio-frequency amplification is obtainable, about 40 to 75 per stage, using the CX-322 as a detector and audio-frequency amplifier either as a screen-grid tube or as a space charge tube. This high audio-frequency amplification is not

* Original Manuscript Received by the Institute, March 28, 1928.

always desirable for several reasons. First, the coupling between tubes, due to common voltage supply circuits, becomes more serious and may lead to frequency distortion. Second, microphonic disturbances in the detector are also more serious than usual because of the high audio-frequency amplification.

To take full advantage of the CX-322 amplification it was considered desirable to investigate the detector action of the screen-grid tube with the elimination of the first audio-frequency amplifier in view.

This imposes the following requirements on the detector tube:

First, it must be able to utilize efficiently radio-frequency input voltages of several volts without overloading.

Second, it must be capable of supplying 20 to 30 volts to the grid of the power tube.

A consideration of grid-leak detection was eliminated for two reasons; first, it could not fulfill the first requirement mentioned above, that is, efficiently utilize several volts radio-frequency input without overloading, and second, this method would add damping to the tuned circuit and reduce the selectivity.

Plate rectification using the CX-322 was found to fulfill both of the above requirements imposed on the detector tube provided that a suitable detector output circuit was designed. Either impedance or resistance coupling could be used, but the values required are larger than for a three-element tube because of the high internal impedance of the CX-322.

In this paper the mathematical analysis of the operation of the CX-322 tube used for plate rectification is presented together with data and curves confirming the theoretical results.

The high radio-frequency amplification required to furnish a signal voltage of several volts to the detector can be obtained by taking full advantage of the amplification available with screen-grid tube in the r. f. stages. Hull[1] has obtained an amplification of 40 per stage with an overall amplification of 2×10^6 at 1000 kc. This amplification is more than ample, under normal receiving conditions, to supply the required voltage to the CX-322 detector. For example, a signal of about 10 microvolts per meter is as small a signal as it is practical to amplify

[1] Hull, "Measurements of High Frequency Amplification with Shielded Grid Pliotrons." *Physical Review*, 27, 4, 439–454; April, 1926.

because of the noise level. As an extreme case, assume that four volts input to the detector is required, (it will be shown later that this value is more than ample). Then with an effective antenna height of four meters, the available signal voltage is 40 microvolts and the r.f. amplification required is 100,000 or only one-twentieth of the total obtained by Hull. Another way of looking at this problem is to assume the first stage of audio has an amplification of about 36. As the response of the CX-322 is proportional to the square of the input voltage, as will be shown later, we will require the square root of 36 or six times the radio-frequency voltage amplification obtained with the average three-element tube set that will satisfactorily operate a power tube.

For a more complete description and theory of the screen grid tube see Hull and Williams.[2] In the theoretical screen-grid tube the mutual conductance is the only factor affecting the output under operating conditions. This is due to the plate current being independent of plate voltage, and also because there is practically no capacity between the control grid and plate, thus reducing all tube capacities to circuit constants. The voltage amplification is equal to the mutual conductance times the external impedance. μ in this tube varies with the plate, control-grid and screen-grid voltages, but has a definite value as soon as all the voltages are specified. The coefficients will be expressed in turns of the mutual conductance whenever it is possible as this is easily found and it simplifies the analysis.

The basis of attack of the problem is the method given by Llewellyn,[3] in an article on the three-element thermionic tube. The same method is applied here to the case of the screen-grid tube. The coefficients are in a different form from those used by Llewellyn so that it will be necessary to repeat some of his work in the first part in order to complete the analysis.

The impedances offered to the carrier and side bands are different and are worked out completely so that the equations could be applied to intermediate frequencies where the carrier and side bands differ by an appreciable amount. The expression for the input voltage squared will contain all the terms so that the best conditions for the modulated second harmonic of the

[2] Hull and Williams, "Characteristics of Shielded Grid Pliotrons." *Physical Review*, 27, 4, 432–438; April, 1926.

[3] F. B. Llewellyn, "Operation of Thermionic Vacuum-Tube Circuits," Bell System *Technical Journal*, V, 3, 433–463; July, 1926.

carrier may be found if it is desirable to tune to this frequency to increase the selectivity or for any other reason. Approximations are made later and the final results are almost in the same form as the results by Chaffee and Browning,[4] who obtained them by making the approximations first, and neglecting the terms dealing with the second harmonics of the carrier and side bands.

Consider the circuit shown in Fig. 1,

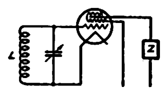

Fig. 1

where Z is a general impedance.

There is no input impedance as there is no mutual capacity between the control grid and plate. The control grid to filament capacity is part of the tuning capacity C. This would not be the case for a three-element tube. There is no low-frequency grid impedance; hence, there can be no grid rectification. The control grid is biased negatively so there is no control grid current, making the impedance from the control grid to filament almost infinite.

The plate current is a function of three variables, E plate, E screen grid, and E control grid with E filament constant. The screen grid is at a fixed positive d-c. potential and as there is no impedance in its circuit, it is at zero a-c. potential, making its effect constant. The plate current will be a function of only two variables, E plate and E control grid.

Eg will refer to the control grid and as the screen grid Ec_2 is left at a fixed potential the analysis can be made as for a three-element tube.

$$Ip = f(Ep, Eg) \text{ with } Ec_2 \text{ constant.} \tag{1}$$

The following notation will be employed:

$$\begin{aligned} Ip &= {}_0Ip + ip \\ Ep &= {}_0Ep + ep \\ Eg &= {}_0Eg + eg \end{aligned} \tag{2}$$

[4] E. L. Chaffee and G. H. Browning, "A Theoretical and Experimental Investigation for Small Signals," Proc. I.R.E., 15, 2; February, 1927.

where the lower case letters denote variations of the normal or d-c. values of voltages and currents distinguished by the zero subscripts preceding the letters.

$$ {}_0Ip+\delta_0 Ip=f({}_0Ep+\delta_0 Eg,{}_0Eg+\delta_0 Eg). \tag{3}$$

Expanding (3) by the extension of Taylor's Theorem we have,

$$ {}_0Ip+\delta_0 Ip=f({}_0Ep{}_0Eg)+\frac{\partial_0 Ip}{\partial Eg}\delta_0 Eg+\frac{\partial_0 Ip}{\partial Ep}\delta_0 Ep+\frac{1}{2}\frac{\partial^2{}_0 Ip}{\partial Eg^2}\delta_0 Eg^2 + \frac{\partial^2{}_0 Ip}{\partial Eg\partial Ep}\delta_0 Ep\delta_0 Eg+\frac{\partial^2{}_0 Ip}{\partial Ep^2}\delta_0 Ep^2+\cdots \tag{4}$$

Which may be written as

$$ip=\frac{\partial_0 Ip}{\partial Eg}eg+\frac{\partial_0 Ip}{\partial Ep}ep+\frac{1}{2}\frac{\partial_0{}^2 Ip}{\partial Eg^2}eg^2+\frac{\partial^2{}_0 Ip}{\partial Eg dEp}epeg + \frac{1}{2}\frac{\partial^2{}_0 Ip}{\partial Ep^2}e^2p+\cdots. \tag{5}$$

Where

$$\frac{\partial_0 Ip}{\partial Eg}=\frac{\mu}{rp} \qquad \frac{\partial^2{}_0 Ip}{\partial Eg\partial Ep}=\frac{\partial gm}{\partial Ep}$$

$$\frac{\partial_0 Ip}{\partial Ep}=\frac{1}{rp} \qquad \frac{\partial^2{}_0 Ip}{\partial Ep^2}=-\frac{1}{rp^2}\frac{\partial rp}{\partial Ep}$$

$$\frac{\partial^2{}_0 Ip}{\partial Eg^2}=\frac{\partial gm}{\partial Eg}$$

and μ = amplification factor
rp = internal plate resistance
gm = mutual conductance.

Eq. (5) is a power series and may be expressed as follows:

$$ip=a_1eg+a_2eg^2+a_3eg^3+\cdots. \tag{6}$$

Before proceeding further it will be necessary to consider complex quantities. First assume:

$$eg=E\cos\omega t. \tag{7}$$

Now by the well-known theory of the complex variable

$$eg=\frac{E\epsilon^{j\omega t}+E\epsilon^{-j\omega t}}{2} \text{ or } \frac{e}{2}+\frac{\bar{e}}{2}. \tag{8}$$

Where the bar over e indicates that it is the conjugate of e.

$$I=\frac{\frac{E}{2}e^{j\omega t}}{Z(j\omega)}+\frac{\frac{E}{2}e^{-j\omega t}}{Z(-j\omega)} \text{ for steady conditions.}^{5} \tag{9}$$

Z for any network may always be expressed as series impedance.

$$Z=r+pL+\frac{1}{pc} \tag{10}$$

For steady conditions $j.\omega.$ may be substituted for p in Eq. (10), this will be done as we are only interested in the steady conditions.

$$Z(j\omega)=r+j\omega L+\frac{1}{j\omega C} \tag{11}$$

$$Z(-j\omega)=r-j\omega L-\frac{1}{j\omega C} \tag{12}$$

$\frac{1}{Z(j\omega)}$ and $\frac{1}{Z(-j\omega)}$ are admittances and may be expressed as:

$$\frac{1}{Z(j\omega)}=a, \text{ and } \frac{1}{Z(-j\omega)}=\bar{a} \tag{13}$$

For detection we will be interested in a modulated radio-frequency voltage of the form

$$eg=A(1+B\cos pt)\cos qt \tag{14}$$

where q is the low frequency
p is the radio frequency
A is the peak value of eg
B is the per cent modulation

$$eg=A\cos pt+\frac{AB}{2}\cos(p+q)t+\frac{AB}{2}\cos(p-q)t. \tag{15}$$

Eg from Eqs. (8) and (15) may be expressed as

$$eg=\frac{A}{2}({}_{k}e_1+{}_{k}\bar{e}_1)+\frac{AB}{4}({}_{k}e_1+{}_{k}\bar{e}_1+{}_{n}e_1+{}_{n}\bar{e}_1). \tag{16}$$

[5] "Electric Circuit Theory and the Operational Calculus," by John R. Carson, page 9.

where the letter preceding refers to the frequency and the number following to the order

and
$$h=p$$
$$k=p+q$$
$$n=p-q$$

Omitting conjugates, eg^2 with its coefficients becomes

$$\left.\begin{aligned}
eg^2 = \frac{A^2B^2}{4}\,{}_{k+n}e_2 + \frac{A^2B^2}{8}\,{}_{2k}e_2 + \frac{A^2B^2}{8}\,{}_{2n}e_2 \qquad & (1)\\
\frac{A^2B}{2}\,{}_{h+k}e_2 + \frac{A^2B}{2}\,{}_{h+n}e_2 + \frac{A^2}{2}\,{}_{2h}e_2 \qquad & (2)\\
\frac{A^2}{2}\,{}_{0h}e_2 + \frac{A^2B^2}{8}\,{}_{0k}e_2 + \frac{A^2B^2}{8}\,{}_{0n}e_2 \qquad & (3)\\
\frac{A^2B}{2}\,{}_{h-k}e_2 + \frac{A^2B}{2}\,{}_{h-n}e_2 + \frac{A^2B^2}{4}\,{}_{k-n}e_2 \qquad & (4)
\end{aligned}\right\} \quad (17)$$

Line 2 represents the second harmonic of the carrier modulated with the low frequency q. Line 1 represents the second harmonics of the carrier and side bands. These second harmonics of the side bands beating against the second harmonics of the carrier would give a frequency $2q$ causing distortion. Thus, if it were desirable to tune to the second harmonic and then to detect again there would be a distortional second harmonic introduced by the first detection.

Line 3 contains the terms causing a change in the direct current of the plate circuit. Line 4 contains the low-frequency modulating frequency q with its second harmonic $2q$, the distortional component.

Eq. (6) becomes, after substituting for eg and eg^2 the values found in Eqs. (16) and (17) and neglecting the coefficients A and B,

$$\begin{aligned}
ip = {}_h a_{1\,h}e_1 + {}_h\bar{a}_{1\,h}\bar{e}_1 + {}_k a_{1\,k}e_1 + {}_k\bar{a}_{1\,k}\bar{e}_1 + {}_n a_{1\,n}e_1 + {}_n\bar{a}_{1\,n}\bar{e}_1 \\
+ \sum_{hkn} {}_h a_{2\,h}e_2 + \sum_{hkn} {}_h\bar{a}_{2\,h}\bar{e}_2 + \sum_{hkn} {}_{0h}a_{20\,h}e_2 \\
+ \sum_{h+k,\,h+n,\,k+n} {}_{h+k}a_{2\,h+k}e_2 + \sum_{h+k,\,h+n,\,k+n} {}_{h+k}\bar{a}_2 + {}_{h+k}\bar{e}_2 \\
+ \sum_{h-k,\,h-n,\,k-n} {}_{h-k}a_{2\,h-k}e_2 + \sum_{h-k,\,h-n,\,k-n} {}_{h-k}\bar{a}_{2\,h-k}\bar{e}_2
\end{aligned} \qquad (18)$$

$$\text{also } ep = -\sum Zip \tag{19}$$

where Z is the impedance for the currents of different frequencies.

We can write Eq. (5) as

$$ip = \frac{\mu}{rp}eg + \frac{1}{rp}ep + \frac{1}{2}eg^2\frac{\partial gm}{\partial Eg} + epeg\frac{\partial gm}{\partial Ep} - \frac{1}{2}\,\frac{1}{rp^2}ep^2\frac{\partial rp}{\partial Ep}. \tag{20}$$

Substitute the value of eg from (16) and the value of eg^2 from (17), neglect the coefficients A and B and equate values of ip from (18) and (20).

Equating the coefficients of e we obtain for the first order effects,

$$\sum_{hkn} {}_h a_1\, {}_h e_1 = \sum_{hkn} \frac{\mu}{rp}\, {}_h e_1 - \sum_{hkn} \frac{1}{rp}\, {}_h a_1\, {}_h e_1$$

or

$${}_h a_1\, {}_h e_1 \left[1 + \frac{{}_h Z_1}{rp}\right] = \frac{\mu}{rp}\, {}_h e_1$$

hence

$${}_h a_1 = \frac{\mu}{rp + {}_h Z_1} \tag{21}$$

${}_k a_1$ and ${}_n a_1$ will be of the same form.

This is the usual amplifier equation and will not be further considered here.

Solving for ${}_{h-k} a_2$

$$\begin{aligned} {}_{h-k}a_2\,{}_{h-k}e_2 = &-\frac{1}{rp}\,{}_{h-k}a_2\,{}_{h-k}e_2\,{}_{h-k}z + \frac{1}{2}\,{}_{h-k}e_2\frac{\partial gm}{\partial Eg} - \frac{\partial gm}{\partial Ep}\,{}_k\bar{z}_{1k}\,\bar{a}_1\,{}_{h-k}e_2 \\ &+ \frac{1}{2}\cdot\frac{1}{rp^2}\,\frac{\partial rp}{\partial Ep}\,{}_h z_{1h} a_{1k}\,\bar{z}_{1k}\,\bar{a}_1\,{}_{h-k}e_2 \end{aligned} \tag{22}$$

or

$$\begin{aligned} 2(h-k)a_2\frac{rp + {}_{h-k}z}{rp} = &\frac{\partial gm}{\partial Eg} - \frac{2\mu_k\,\bar{z}_1}{rp + {}_k\bar{z}_1} \\ &+ \frac{\left(\dfrac{\mu}{rp}\right)^2 {}_h z_{1k}\,\bar{z}_1}{(rp + {}_h z_1)(rp + {}_k\bar{z}_1)}\,\frac{\partial rp}{\partial Ep} \end{aligned} \tag{23}$$

hence

$${}_{h-k}a_2 = \frac{1}{2}\,\frac{rp}{rp + {}_{h-k}z}\left\{\frac{\partial gm}{\partial Eg} - \frac{2\mu_k\,\bar{z}_1}{rp + {}_k\bar{z}_1}\,\frac{\partial gm}{\partial Ep}\right.$$

$$+\frac{\left(\frac{\mu}{rp}\right)^2 {}_h z_1\, {}_k\bar{z}_1}{(rp+{}_h z_1)(rp+{}_k\bar{z}_1)}\frac{\partial rp}{\partial Ep}\Bigg\}$$

similarly

$$ {}_{h-n}a_2=\frac{1}{2}\,\frac{rp}{rp+{}_{h-n}z}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu\,{}_n\bar{z}_1}{rp+{}_n\bar{z}_1}\,\frac{\partial gm}{\partial Ep}\right. \tag{25}$$

$$\left.+\frac{\left(\frac{\mu}{rp}\right)^2 {}_k z_1\, {}_n\bar{z}_1}{(rp+{}_h z_1)(rp+{}_n\bar{z}_1)}\,\frac{\partial rp}{\partial Ep}\right.$$

$$ {}_{k-n}a_2=\frac{1}{2}\,\frac{rp}{rp+{}_{k-n}z}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu\,{}_n\bar{z}_1}{rp+{}_n\bar{z}_1}\,\frac{\partial gm}{\partial Ep}\right. \tag{26}$$

$$\left.+\frac{\left(\frac{\mu}{rp}\right)^2 {}_k z_1\, {}_n z_1}{(rp+{}_n z_1)(rp+{}_n\bar{z}_1)}\,\frac{\partial rp}{\partial Ep}\right.$$

Solving for a_2 of the form, ${}_{0h}a_2$

$$ {}_{0h}e_2\, {}_{0h}a_2=-\frac{1}{rp}\,{}_{0h}a_2\, {}_{0h}e_2\, {}_{0h}z+\frac{{}_{0h}e_2}{2}\,\frac{\partial gm}{\partial Eg} \tag{27}$$

$$-\frac{\partial gm}{\partial Ep}\left\{{}_h\bar{z}_1\, {}_h a_1\, {}_{0h}e_2+\frac{1}{2}\,\frac{1}{rp^2}\,{}_h z_1\, {}_h a_1\, {}_h\bar{z}_1\, {}_h\bar{a}_1\,\frac{\partial rp}{\partial Ep}\right\}$$

$$ {}_{0h}a_2=\frac{1}{2}\,\frac{rp}{rp+{}_{0h}z}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu\,{}_h\bar{z}_1}{rp+{}_h z_1}\,\frac{\partial gm}{\partial Ep}\right.$$

$$\left.+\frac{\left(\frac{\mu}{rp}\right)^2 {}_h z_1\, {}_h\bar{z}_1}{(rp+{}_h z_1)(rp+{}_h\bar{z}_1)}\,\frac{\partial rp}{\partial Ep}\right\} \tag{28}$$

${}_{0k}a_2$ and ${}_{0n}a_2$ are of similar form.

The coefficients for the different second harmonics may be found by the same method.

In most cases

$$pz=p+qz=p-qz \text{ (very nearly)}$$

Then

$$
{}_{h-n}a_2=\frac{1}{2}\,\frac{rp}{rp+qz}\left\{\frac{\partial gm}{\partial Eq}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+pz)}\,\frac{\partial rp}{\partial E}\right\} \tag{29}
$$

and

${}_{h-n}a_2={}_{h-k}\,a_2$ and ${}_{n-k}\,a_2={}_{h-k}\,a_2$ except that $2qz$ in ${}_{n-k}\,a_2$ replaces

qz in ${}_{h-k}\,a_2$. (30)

Also,

$$
{}_{0h}a_2=\frac{1}{2}\,\frac{rp}{rp+R}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+pz)}\,\frac{\partial rp}{\partial Ep}\right. \tag{31}
$$

where R is the d-c. resistance of Z.

$$
{}_{0k}a_2={}_{0n}a_2={}_{0h}a_2 \tag{32}
$$

$$
{}_{q}\dot{i}_{p}=\frac{rp}{rp+qz}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzq\bar{z}}{(rp+pz)(rp+p\bar{z})}\,\frac{\partial rp}{\partial Ep}\right\}\frac{A^2B}{2} \tag{33}
$$

and

$$
{}_{2q}\dot{i}_{p}=\frac{1}{2}\,\frac{rp}{rp+2qz}\left\{\frac{\partial gm}{\partial Eg}-\frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep}\right.
$$

$$
\left.+\frac{\left(\dfrac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+p\bar{z})}\,\frac{\partial rp}{\partial Ep}\right\}\frac{A^2B^2}{4} \tag{34}
$$

$$
{}_0 i_p = \frac{rp}{rp+R}\left\{\frac{\partial gm}{\partial Eg} - \frac{2\mu pz}{rp+pz}\,\frac{\partial gm}{\partial Ep} + \frac{\left(\frac{\mu}{rp}\right)^2 pzp\bar{z}}{(rp+pz)(rp+p\bar{z})}\,\frac{\partial rp}{\partial Ep}\right\}\left\{1+\frac{B^2}{2}\right\}\frac{A^2}{4}. \tag{35}
$$

The factor $\frac{\partial gm}{\partial Eg}$ may be found graphically by plotting gm against Ec_1 and drawing tangents. Typical static curves for $\frac{\partial gm}{\partial Eg}$ are shown in Figs. 2 and 3. The dynamic curves will be dis-

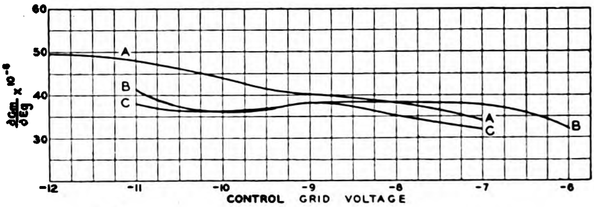

Fig. 2—CX-322; $\frac{\partial Gm}{\partial Eg}$ Control Grid Voltage Characteristics.
A—Static Value $Eb=135$, $Ec_2=67.5$
B—Dynamic Value $Eb=225$, $Ec_2=67.5$
C—Dynamic Value $Eb=135$, $Ec_2=67.5$

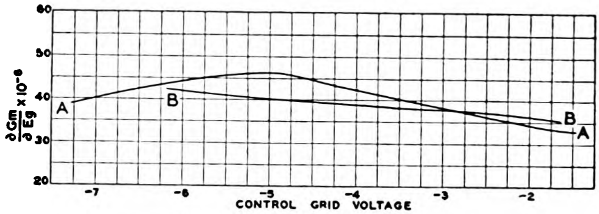

Fig. 3—CX-322. $\frac{\partial Gm}{\partial Eg}$—Control Grid Voltage Characteristics.
A—Static Value $Ec_2=+45$, $Eb=135$
B—Dynamic Value $Ec_2=+45$, $Eb=135$

cussed later. The noticeable feature about these curves is that under the d-c. conditions given $\frac{\partial gm}{\partial Eg}$ remains practically constant. It was found that if Ec_2 was varied between about 30 and 67.5 and Ec_1 adjusted to give a plate current of approximately 100 microamperes with no signal voltage, the values of $\frac{\partial gm}{\partial Eg}$

were practically constant but increased slightly as the screen-grid voltage was lowered. The region where Ec_2 was under 30 volts was not investigated because it was desired to keep the control grid bias at least 7 volts negative.

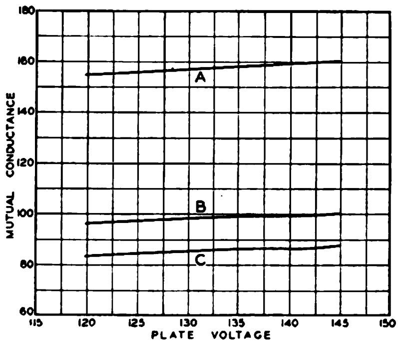

Fig. 4—Mutual Conductance—Plate Voltage Characteristics of CX-322.
A—$Ec_1 = -1.5$, $Ec_2 = 22.5$
B—$Ec_1 = -7.5$, $Ec_2 = 45.0$
C—$Ec_1 = -12.0$, $Ec_2 = 67.5$

The next two terms may be neglected as their values are small. Fig. 4. shows gm plotted against Eb. The resultant slope is very small so $\dfrac{\partial gm}{\partial Eg}$ is very nearly zero. The factor μ is large but

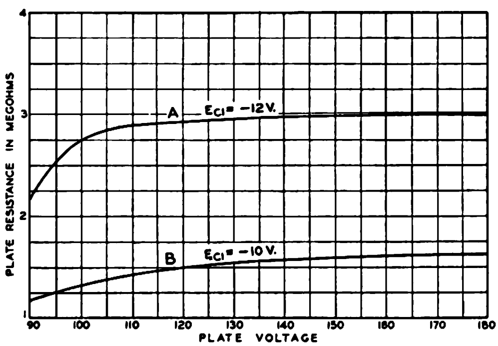

Fig. 5—Plate Resistance—Plate Voltage Characteristics of CX-322.
A—$Ec_1 = -12.0$, $Ec_2 = +67.5$
B—$Ec_1 = -10.0$, $Ec_2 = +67.5$

rp is very large so that the factor $\dfrac{\partial gm}{\partial Ep}$ is small under detection conditions making the effect of the first term negligible. The value of rp plotted against Eb is shown by Fig. 5. This variation

Refer to Eq. (35). If R is made zero ${}_pZ$ and its conjugate ${}_p\bar{z}$ are also zero and Eq. 35 becomes

$$ {}_0ip = \frac{\partial qm}{\partial Eg}\left(1+\frac{B^2}{2}\right)\frac{A^2}{4}. \qquad (36)$$

If a known input voltage of any frequency is introduced on the grid the value of $\frac{\partial gm}{\partial Eg}$ can be determined from the direct current change if B is zero or if its value is known. The dynamic values of $\frac{\partial gm}{\partial Eg}$ were found by this method. The agreement between the static and dynamic values is fair considering the difficulties of drawing accurate tangents to curves and the number of times experimental errors enter into the results. As far as the author knows this method of obtaining $\frac{\partial gm}{\partial Eg}$ is a new one.

Fig. 6 shows the plate-current change plotted against plate voltage. The full lines are for 60 cycles and the dotted lines are for 1000 kc. The reason the lines are displaced is that two different tubes were used. The lines are approximately parallel. A condenser was placed across the control grid to filament for the 60 cycles input so as to offer the same impedance as the tube capacity would have at 1000 kc. These curves prove that the value of $\frac{\partial gm}{\partial Eg}$ is very nearly constant for any input voltage nearly as great as the control grid bias under proper conditions, but falls off slightly as the input voltage is increased.

Fig. 7 shows the plate current plotted against plate voltage with different load lines drawn in from 270 volts. The output voltage for a given input voltage and load may be obtained from these curves. For example, with 400,000 ohms load and 2 volts r.m.s. input modulated 50 per cent the output voltage would be 52 volts. This is obtained by taking point A as a reference. The a-c. voltage across the resistor for the above conditions would vary between B and C so the output voltage would be the difference, approximately 52 volts. This is the total swing so the peak value is one half of this, or 26 volts. For 40 per cent modulation the peak value would be eighty

Fig. 8 shows the d-c. plate current plotted against Eb with 400,000 ohms in the plate circuit. The values of pZ and $p\overline{Z}$ are not zero here so if the last two terms of the coefficients of Eq. 35 are not small the d-c. should change when the external impedance is changed. The resistor was shorted for r.f. voltages by a condenser, but no appreciable change could be seen in the direct current.

Eqs. (33), (34), and (35) become

$$_q\dot{i}_p = \frac{rp}{rp+qz} \frac{\partial gm}{\partial Eg} \frac{A^2B}{z}. \tag{37}$$

$$_{2q}\dot{i}_p = \frac{rp}{rp+2qz} \frac{\partial qm}{\partial Eg} \left\{\frac{A^2B^2}{4}\right\}. \tag{38}$$

$$_0\dot{i}_p = \frac{rp}{rp+R} \frac{\partial gm}{\partial Eg} \left\{1+\frac{B^2}{2}\right\} \frac{A^2}{4}. \tag{39}$$

Fig. 8 departs from the square law for two reasons; first, a decrease of $\frac{\partial gm}{\partial Eg}$ with input voltage, and second, rp decreases with input voltage. If R remains constant the value of $\frac{rp}{rp+R}$ will decrease with a decrease of rp. Fig. 9 shows value of rp plotted against Ib with Ec_2 and Eb remaining constant. If R is 400,000 ohms and rp varies from four megohms to one megohm the value of the fraction $\frac{rp}{rp+R}$ varies from 0.91 to 0.71. These would be extreme variations. Some variation from the square law is expected from the theory taking everything into account.

In order to determine the value of the input voltage required to work a power tube an example will be worked out. Resistance coupling will be used to couple the plate of the detector to the power tube. The circuit used is shown by Fig. 10.

The following values were used:

$$R = qz = 2qz = 5 \times 10^5 \text{ ohms}$$

$$R \text{ grid } = 2 \times 10^6 \text{ ohms}$$

$$C \text{ coupling} = 0.2 \times 10^{-6} \text{ farads}$$

The a-c. impedance of the combination is practically 4×10^5 ohms. Fig. 8 is plotted with an impedance of 4×10^5 ohms.

When B is zero the change in d-c. determines the value of $\frac{rp}{rp+R}\,\frac{\partial gm}{\partial Eg}\,\frac{A^2}{4}$ as can be seen by referring to Eq. (30). This is one-half of the value of the coefficient of B in Eq. (36). The current qip will be two times the change in d-c. found from curve

Fig. 9—Plate-Current, Plate-Resistance Characteristics of CX-322.

8 times B. If this is multiplied by the external impedance we obtain the voltage ${}_qe_p$, that is, applied to the grid of the next tube.

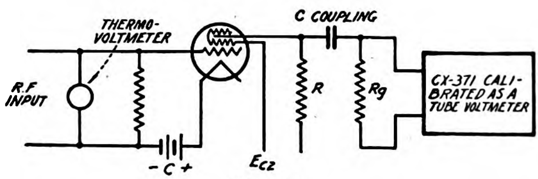

Fig. 10

Assume $B=0.35$ and $A=2\sqrt{2}$.

$$ {}_qe_p=2\times B\times 64\times 10^{-6}\times 4\times 10^5=17.92 \text{ volts}$$

$$ {}_{2q}e_p=B^2\times 64\times 4\times 10^5\times 10^{-6}=3.07 \text{ volts}$$

For $B=0.35$ and $A=\sqrt{2}$

$$ {}_qe_p=2\times B\times 17\times 10^{-6}\times 4\times 10^5=4.26 \text{ volts}$$

$$ {}_{2q}e_p=B^2\times 17\times 10^{-6}\times 4\times 10^5=0.81 \text{ volts}$$

To verify these values a CX-371 was calibrated as a tube voltmeter and the detector was worked directly into it as

shown in Fig. 10. The same values of impedances were used as in the calculations. An r.f. choke was placed in series with the coupling condenser so that with no modulation the voltmeter read zero. When the carrier was modulated the tube voltmeter read the peak values of the output voltage which included ${}_qe_p$ and $2{}_qe_p$. The results for B equal to forty per cent are given in Table I.

TABLE I
INPUT VOLTAGE TO CX-371 TUBE VOLTMETER
B = 40 per cent

A	$\sqrt{2}$	$2\sqrt{2}$
Tube Voltmeter Reading	6.0	20.8
Calculated Voltage	5.57	21.0

The external resistor was shorted with a condenser so as to change ${}_pZ$ and ${}_p\overline{Z}$. In both cases the voltage decreased very slightly. This again proves that the last two terms of the coefficients of A^2 in Eqs. (33), (34), and (35) are very small and may be neglected in the analysis. This could be expected from the considerations of the characteristics of the CX-322. The internal impedance is very high. The tube has some plate to filament capacity and this with the wiring capacity has a lower impedance than the internal impedance at radio frequencies. This impedance is in parallel with the load, so the resultant external impedance would be much less than the internal impedance. There would be very little gained by shorting the external impedance for radio frequencies when the tube is used as a detector.

If Eq. (33) is multiplied by $rp+qZ_1$, Eq. (34) by $rp+2qZ_1$, and (35) by $rp+R$ we get the equivalent voltages introduced in the plate circuit. These voltages may be used as μeg in amplifier equations. In order to calculate the voltage on the grid of the next tube it is only necessary to solve for the voltage across the grid using the above equivalent voltages as in the regular amplifier equations.

In conclusion the writer wishes to thank Mr. R. M. Wise and Mr. D. F. Schmit for their helpful comments and suggestions in the preparation of this paper.

THE SCREEN-GRID TUBE*

By

N. H. Williams

(University of Michigan, Ann Arbor, Michigan)

Summary—*Radio-frequency amplification by means of the three-electrode tube is usually disappointing. With resistance coupling the feed-back through the tube is in such a phase relation to the input voltage as to reduce the amplification below that given by the simple equation for voltage amplification.*

With impedance coupling the feed-back causes self-oscillations when the circuit conditions are such as might be expected to give large amplification.

In the shielded grid tube the feed-back is reduced to a negligible amount and the current through the tube is very nearly independent of the plate voltage over the working range. Under these conditions, the voltage amplification becomes the product of the mutual conductance and the load impedance. High impedance in the plate circuit is obtained by using a sharply tuned parallel circuit. With proper shielding such a circuit may be used without producing self-oscillation. At a frequency of 700,000 cycles per second, amplifications of 80 fold per stage may be obtained.

IN the case of resistance-coupled amplifiers, we may write the expression for the voltage amplification of one stage in the form $\mu R/R_0+R$, in which μ is the amplifying factor of the tube, R the external or load resistance, and R_0 the resistance of the tube. In such a system the plate of one tube is connected through a grid condenser to the grid of the next tube. This expression for voltage amplification is based upon the assumption of ideal conditions which are not accurately realized. There are two reasons why this amplification cannot be attained. First: as the plate potential varies through a large amplitude, it induces a potential on the grid which is in phase with itself. But the plate potential differs from the potential of the positive battery terminal by the amount of the RI drop in the load resistance and is therefore at its minimum value when the current through the tube is a maximum. This results in a feed-back through the tube to the grid in opposite phase to the actual grid voltage, thus reducing the effectiveness of the grid in controlling the current. Second: The equation indicates that the amplification may be increased by increasing the load-resistance indefinitely, the limit being the μ of the tube. As a matter of fact, the capacitance from

* Original Manuscript Received by the Institute, January 20, 1928. Delivered before the Detroit Section of the Institute. December 16 1927.

plate to filament in the tube is in parallel with the load resistance and hence the impedance of the circuit is increased very little by adding resistance beyond a certain limit. It is for these reasons that resistance coupling for high-frequency amplification is usually disappointing.

With impedance coupling by means of a tuned circuit between plate and battery instead of a pure resistance, the case is very different. The tuned circuit is adjusted to behave as an inductive reactance and then the feed-back through the tube is so changed in phase that regeneration takes place and the amplification is greater than would be expected from the equation that is used in representing it. Furthermore, the capacitance from plate to filament is simply added to that of the condenser in the tuned circuit and is "tuned out"; thus no reduction of impedance results. The feed-back through the tube is so effective in this case as an agent of regeneration that the system breaks into self-oscillations and becomes useless as an amplifier if a sharply tuned circuit is employed to obtain impedance coupling.

The screen grid is a device to prevent the feed-back from plate to grid and thus prevent self-oscillation when the plate circuit is highly tuned. The second grid acts as an electrostatic screen between the plate and the control grid. It is held at a fixed potential and cuts off the control grid from the influence of the fluctuating potential of the plate. This tube for radio-frequency amplification was developed at the Research Laboratory of the General Electric Co. by A. W. Hull and N. H. Williams.[1]

Measurements show that the capacitance from plate to control grid is reduced to 1 per cent of that in the three-electrode tube when the latter is used without its socket. The percentage is much lower when the screen-grid tube is compared with the three-electrode tube as normally used. With these tubes it is possible to use sharply tuned circuits between plate and battery at almost any frequency without causing self oscillations. An amplification of 60 fold per stage is easily possible at a frequency of 700,000 cycles per second, and at lower frequencies still higher amplifications may be reached. An over-all amplification of more than two million-fold has been measured for a five-stage amplifier so built that each stage was in a separate compartment of a metal box.

[1] *Physical Review*, April, 1926.

The tube has a mutual conductance of about 0.4 milliampere per grid volt. Its plate-voltage, plate-current characteristic shows a downward slope where the plate voltage is less than the potential of the screen grid. Thus the tube has negative resistance in this region. This effect is the result of secondary emission from the plate. Each electron that strikes the plate may dislodge several electrons from the atoms of the metal, in which case the current in the plate circuit is reversed. When the plate voltage is above that of the screen grid, the current is positive, the characteristic is nearly straight and nearly parallel to the voltage axis. Here the tube resistance to alternating impulses is very high, which signifies that the plate current is practically independent of the plate voltage. This is an important item in connection with the behavior of the tube, for it accounts for an amplifying factor of over 200 and it eliminates one of the variables which, with other tubes, must be considered in connection with the circuit.

When the three-electrode tube is used in a radio circuit, the high-frequency component of the plate current is a function of seven different properties of the tube. These are:

(1) Rate of change of plate current with grid voltage.
(2) Rate of change of plate current with plate voltage.
(3) Rate of change of grid current with grid voltage.
(4) Rate of change of grid current with plate voltage.
(5) Capacitance between plate and grid.
(6) Capacitance between plate and filament.
(7) Capacitance between grid and filament.

As explained above, the second of this list is eliminated when the screen-grid tube is used because of the flatness of the characteristic curve. When the tube is used as an amplifier, the grid current may be kept at zero value by proper grid bias, and thus the third and fourth items are disposed of. In the screen-grid tube the capacitance from plate to grid and from plate to filament is negligible. The capacitance from grid to filament becomes a circuit constant since it is in parallel with another condenser. There remains, then, only the variation of plate current with grid voltage, i.e., mutual conductance G, and it appears that when this tube is used the mathematical solution of circuits is greatly simplified.

The amplifying factor, μ, of the tube expresses how many times more effective an increment of voltage is if applied to the

grid than if applied in the plate circuit. In mathematical symbols it is represented by the ratio of two differential coefficients.

$$\mu = \partial I_p/\partial E_g \div \partial I_p/\partial E_p$$
$$\partial I_p/\partial E_g = \quad G \quad = \text{Mutual conductance.}$$
$$\partial I_p/\partial E_p = 1/R_0 = \text{Plate conductance.}$$
$$R_0 = \text{tube resistance.}$$

From this it follows that $\mu = GR_0$.

Substituting GR_0 for μ in the equation with which we started, we have voltage amplification $= GR_0R/R+R_0$.

If R_0 is large as compared with R, we may neglect R in the denominator and the expression for voltage amplification reduces to GR. In the case of impedance coupling, if Z represents the impedance, it is apparent that GZ becomes an approximate expression for the voltage amplification. With a parallel-tuned circuit, the impedance is given by the equation

$$Z = L/CR$$

L being the inductance in the circuit, C the capacitance, and R the resistance. It is not difficult in a carefully tuned circuit to raise the impedance to 150,000 ohms and if G is taken as 0.4 milliampere per grid volt, we have

$$\text{Voltage amplification} = GZ = 60.$$

This emphasizes the simplicity of the computations in the screen-grid amplifier.

The first piece of research that was undertaken making use of the shielded grid amplifier was the measurement of the charge of the electron by the shot effect.[2] This was done with an amplifier that was constructed at the Research Laboratory of the General Electric Co. Because of the fact that electricity is not an infinitely fine-grained fluid, but consists of discrete particles or electrons, the flow of the charge to the plate in a vacuum tube causes minute fluctuations of the plate potential due to the random distribution of the electrons in time. These minute fluctuations of voltage were measured with the screen-grid amplifier and from these measurements the charge of the electron was computed. The results agree within one per cent with those obtained by other and totally independent methods. Electrons produced by thermionic emission, by photoelectric emission, and by ionization of gases were measured in this way and now the measurement of the positive charge of ions is going on at the University of Michigan.

[2] Hull and Williams, *Physical Review*, February, 1925.

Discussion on
THE MEASUREMENT OF CHOKE COIL INDUCTANCE*
(C. A. Wright and F. T. Bowditch)

W. O. Osbon†: The authors have given an interesting and valuable discussion of several simple methods of measuring, under operating conditions, the inductances of choke coils carrying direct current. They have made one or two statements, however, which the present writer desires to comment upon.

The inductance of a coil with a closed iron core is given by

$$L = \frac{0.4\pi N^2 \mu A}{l} \times 10^{-8}$$

where N = number of turns in winding
A = area of core section, sq. cm.
l = length of iron path, cm.
μ = effective permeability of the iron corresponding to conditions under which the choke is operating.

It is seen that all the factors entering into the above equation except the permeability are determined by the geometry of the coil, the permeability alone being affected by operating conditions. Hence, an investigation of the variations of μ under different conditions will lead to an understanding of the variations in inductance for a given coil.

Concerning the effective permeability of a core with an a-c. magnetizing force superposed on a d-c. mmf. the authors have made a serious error. They have stated that the inductance is "determined by the average slope of the saturation curve over the range within which the current varies," and have neglected hysteresis entirely. When an a-c. mmf. of maximum value $1/2\,\Delta H$ is superposed on the d-c. mmf. there will be produced a minor hysteresis loop[1] as at a in the accompanying sketch. The total change in mmf., ΔH, will produce a change in density, ΔB, and the effective or incremental permeability[1] is

* Proc. I.R.E., 16, 3; March, 1928.

† Research Laboratory, Westinghouse Elec. & Mfg. Co., East Pittsburgh, Pa.

[1] T. Spooner, "Permeability" *Jour.* A.I.E.E. Vol. XLII, p. 42 and T. Spooner "Effect of Superposed Alternating Field on Apparent Magnetic Permeability and Hysteresis Loss", *Phys. Rev.* 1925.

$$\mu_\Delta = \frac{\Delta B}{\Delta H},$$

or it is the slope of the line drawn through the tips of the minor loop, which is seen to be quite different from the average slope of the normal curve in the range of variation of H. The minor loops at b and c indicate the state of affairs for values of d-c. mmf. which nearly saturate the core. It is seen that the incremental permeability decreases for increasing values of d-c. mmf. The writer has made tests on several kinds of iron which show that this statement holds even for very low values of d-c. magnetization in the range of upward curvature of the magneti-

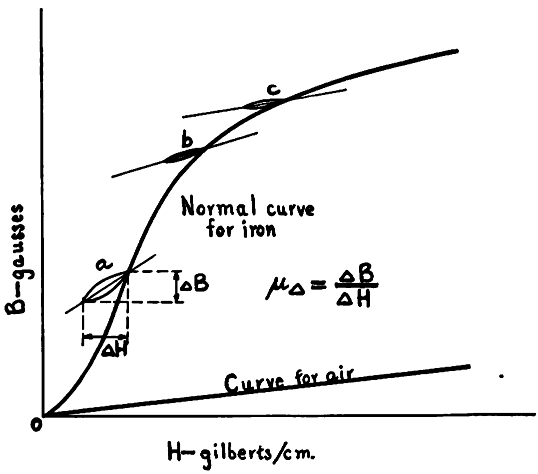

Fig. 1

zation curve. It seems likely to the writer, therfore, that the drop in inductance with no direct current shown in Fig. 6 in the paper was due to the presence of some residual magnetism in the core.

The value μ_Δ corresponding to the particular values of a-c. and d-c. magnetization should be substituted for μ in the formula given above. In the second paper referred to above, Mr. Spooner gives a formula for calculating μ_Δ. When there is an air gap of length a in the magnetic circuit the formula for inductance takes the form

$$L = \frac{0.4\pi N^2 A}{\dfrac{l}{\mu_\Delta} + a} \times 10^{-8}$$

In a paper by C. R. Hanna,[2] a direct method is developed for calculating the best value of air gap to be used in the magnetic circuit of a choke coil carrying direct current. In a recent paper in QST,[3] curves are given for calculating chokes according to the method developed by Hanna, for several different kinds of iron not considered in Hanna's paper.

In the latter part of their paper, the authors commit a serious error by stating that the slope of the magnetization curve for air is considerably greater than the slope of the magnetization curve for iron in the region of saturation. The truth of the matter is that the slope of the iron curve *approaches* the slope of the curve for air. In other words, the incremental permeability of iron is never less than unity as the statement of the authors would lead one to believe.

C. A. Wright†: The authors are familiar with the minor loops referred to by Mr. Osbon and have referred to them in a previous discussion.[1] However, it was decided that in the present case they were of negligible importance and that a consideration of them would unnecessarily complicate the methods of inductance measurement suggested.

The present paper was intended to apply to choke coils used in radio "B" power units, the inductance of which was the subject of much interest at the time the measurements were made.

In the cases of a representative lot of commercial samples measured the inductances were practically the same for increasing and decreasing values of direct current. With the negligible hysteresis thus indicated, the slopes of the minor loops approached closely the slopes of the saturation curve corresponding to the positions of the minor loops, and under these conditions the statements made regarding the value of inductance are correct.

The criticism of Mr. Osbon, that "the slope of the magnetization curve for air is considerably greater than the slope of the curve for iron in the region of saturation," is probably based on the assumption that the abscissas of the curves are measured in

[2] C. R. Hanna, "Design of Reactances and Transformers which Carry Direct Current," *Jour.* A.I.E.E., Feb. 1927, page 128.

[3] D. E. Replogle, "Notes on the Design of Iron-Core Reactances which Carry Direct Current," QST, Apr. 1928, page 23.

† National Carbon Co., Inc., Cleveland, Ohio.

[1] "Telephone Communication," by Wright and Puchstein, Chap. 7, pg. 141–142. Published by McGraw-Hill Book Company, 1925.

the usual magnetizing force per unit length. However, it is to be noted that these abscissas are in "ampere turns" so that the curve for the iron referred to is that of the total length of the iron magnetic circuit, and the curve for air is that of the total length of the air gap. If this air gap is very small, as in the actual case, this curve for air may be more steep, as is stated in the discussion, than that for the curve of the iron portion of the circuit. The "proportional addition" of the curves for iron and air will then have the stated effect on the total curve.

Discussion on
ON THE DISTORTIONLESS RECEPTION OF A MODULATED WAVE AND ITS RELATION TO SELECTIVITY
(F. K. Vreeland)*

V. D. Landon†: The paper by Frederick K. Vreeland, "On the Distortionless Reception of a Modulated Wave and its Relation to Selectivity," was read with interest at the Westinghouse Electric and Manufacturing Company laboratories. A good deal of work has been done here on band-pass filters somewhat similar in design to that described by Dr. Vreeland. It seems desirable to amplify his discussion considerably.

Of course, it will be realized that the scheme outlined is by no means new either with identical or spaced tuning. The circuit shown is merely two conductively coupled tuned circuits with critical coupling or slightly greater than critical coupling. The same principle has been incorporated in commercial receivers for years, one such use being in a superheterodyne provided with intermediate frequency transformers having tuned primary and tuned secondary windings, the coupling between which is quite loose, being only slightly greater than critical coupling. This, of course, gives the well-known response curve characteristic of such an arrangement in accordance with well understood principles. A treatment of coupled circuits that describes the phenomena very well is given in Morecrofts "Principles of Radio Communication" on page 100. The dotted curve of Fig. 99 illustrates the band pass action perfectly. The precise method of coupling employed is of course immaterial.

The idea of tuning both circuits over a considerable frequency band is not novel either. The old "loose couplers" and similar devices were in use before radio broadcasting was known. These receivers employed tuned primary and tuned secondary, and the coupling was adjustable. In general, something approximating critical coupling was striven for, as this gave the best sensitivity retaining good selectivity.

*Presented at the Annual Convention of the Institute of Radio Engineers, January 9, 1928. Proc. I.R.E., 16, 3; March, 1928.

† Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania.

It is of interest to note the behavior of two such coupled circuits when they are tuned to various frequencies throughout the broadcast range. It is well-known that for critical coupling the mutual reactance should equal the square root of the product of the primary and secondary resistances. Or for duplicate circuits the mutual reactance should equal the resistance of one of the circuits. If greater coupling than this is used two peaks will occur in the resonance curve, the greater the coupling the more pronounced the peaks. For coupling only slightly greater than critical a nearly flat top curve, with steep sides, is obtained. For less than critical coupling the curve resembles an ordinary resonance curve, though a more favorable ratio will be obtained of the width near the top to the width near the bottom. A value of coupling just greater than critical gives the best shape to the resonance curve.

It is a well-known fact that the resistance of a coil varies rapidly with frequency. Common types of coils used in broadcast receiver design vary in resistance at a rate lying between the first and second power of the frequency. If a mutual inductive reactance is used to couple the circuits, this reactance will vary in value exactly as the first power of the frequency. If coils could be found whose resistance varied exactly as the first power of the frequency then the value of coupling could be made correct to give the flat top curve all over the range. The fact that the resistance varies faster than the frequency means that the curve will either have a rounded top at high frequencies (corresponding to less than critical coupling) or a fairly pronounced double peak at low frequencies (corresponding to more than critical coupling).

It is to be noticed, however, that even assuming that the resistance varies in the correct way to give the same curve shape at all frequencies, the width of the curve will not remain constant over the range. The width will be directly proportional to the frequency. That is, the band received will be two and one-half times as wide at the high-frequency end of the scale as it is at the low-frequency end.

Also if the resistance varies as suggested, the signal will be amplified a great deal more at the high frequencies than at the low, since the tuned impedance of the circuit will be directly proportional to the frequency.

If the rate of change of resistance lies between the first and second power of the frequency the selectivity variation over the

range will be even worse than that indicated. The amplification will be more nearly uniform but will still be greatest at high frequencies.

For these reasons it is highly desirable that the coupling increase at a rate slightly less than as the first power of the frequency. This results in decreasing the amplification at high frequencies to the same value that is obtained at low. At the same time the selectivity is made considerably better. Of course, the desirable square top feature is sacrificed to a certain extent but this does not matter as the curve is already far too broad at this end of the scale.

It was found possible to obtain the desired variation in coupling with frequency by employing combinations of capacitive and inductive coupling in a variety of ways.

If this variation is correctly done the resonance curve will be found only slightly wider at high frequencies than at low. At low frequencies the curve had a very slight double peak and at mid-scale an ideal flat top. At high frequencies the curve was quite rounded on top though not nearly so sharp-pointed and sloping-sided as the curve of the same circuits not coupled, operated in cascade.

It is believed that such an arrangement is more nearly ideal than that discussed by Dr. Vreeland.

BOOK REVIEW

Experimental Radio, BY R. R. RAMSEY. Ramsey Publishing Company, Bloomington, Indiana. Third Edition. 229 pages. $2.75.

This compendium of radio experiments was first issued by the author for use in his radio course at Indiana University, where he is Professor of Physics. The first edition appeared in 1923 after about five years of collection of experiments from various sources, including textbooks on radio telegraphy and telephony. The author states that the course is intended to be about on a level with a good second-year college course in physics. No attempt is made to cover the more complicated measurements, but some references to these are included. In all there are 117 experiments. Typical ones are as follows: "To Measure the Dielectric Constant of Oil"; "To Calibrate a Wavemeter Using Overtones"; "Measurement of the Amplification Constant of a Tube"; "To Measure the Resistance of a Radio Circuit; Resistance Variation Method"; "Field Intensity Measurements; Radiation from a Coil."

Within its scope, which the author has clearly defined, the book is an excellent manual of radio- and audio-frequency measurements and demonstrations. While all the material in it should be familiar to radio and electrical engineering graduates, experienced engineers will find Ramsey's outline useful for refreshing their memories on specific points with which their work has not brought them into contact for some time. The only general criticism of the book which might be made is to point out the absurd use of the terms "AC current," "DC current," etc., which are employed with a consistency worthy of a more logical nomenclature. CARL DREHER†

† Staff Engineer, National Broadcasting Company, New York City

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New York	New York City, 40 Rector Street, Room 618	Chesterton, A. J.
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	Washington, c/o National Radio Institute	Sprayberry, Frank L.
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	Chicago, 100 W. Monroe St., c/o R. C. A.	Rothenberger, William
	Chicago, 300 South Racine Avenue	Van Benschoten, Walter
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	Quebec, Lachine, 122 George V Avenue	Boyd, David
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	London, N. W. 6, 15 Lyncroft Gardens	Hinderlich, Albert
	London, W. C. 2, 34 Kingsway	Holzman, Louis
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	Sendai, Kita-samban-cho, No. 76	Saito, Masao
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